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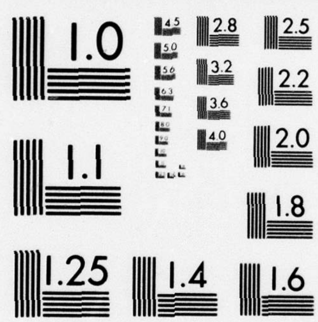

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RESEARCH AND DEVELOPMENT TECHNICAL REPORT

REPORT CORADCOM-76-C-0193

VHF DATA MODEM/BASEBAND DATA SIGNAL CONVERTER

SIGNATRON, INC.
12 HARTWELL AVENUE
LEXINGTON, MASSACHUSETTS 02173

MAY 1978
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VHF DATA MODEM/BASEBAND DATA SIGNAL CONVERTER

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A full duplex data modem, used for transmitting digital data over tactical VHF FM radios, has been designed and built. The modem has two operating modes. In the quasi-analog mode it can transmit at 75, 150, 300, 600, 1200, 2400, and 4800 bps, using binary FSK, MSK, and four-level FSK. In the X-mode it can transmit at 300, 600, 1200, 2400, 4800, 9600, and 16000 bps, using Manchester, conditioned, diphase, pair-selected ternary, delay next page

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modulation, bipolar, and twinned binary codes. This report contains the theory of operation, implementation, and test data on two units.

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PREFACE

This Final Report describes work performed between September 1976 and April 1978 by SIGNATRON, Incorporated, 12 Hartwell Avenue, Lexington, Massachusetts, under Contract DAAB07-76-C-0193, for U.S. Army Communications Research and Development Command (CORADCOM), Fort Monmouth, New Jersey. It documents the VHF Data Modem/Baseband Data Signal Converter, and includes test data.

Mr. Paul Ulrich, DRDCO-COM-RF-2, was the ECOM Project Engineer. Dr. Ehrman was Project Manager. Mr. P. Mahoney was the designer, and Mr. M. Michalik implemented the system. Dr. R. Westerberg performed the analysis of the signal spectra and error rates.

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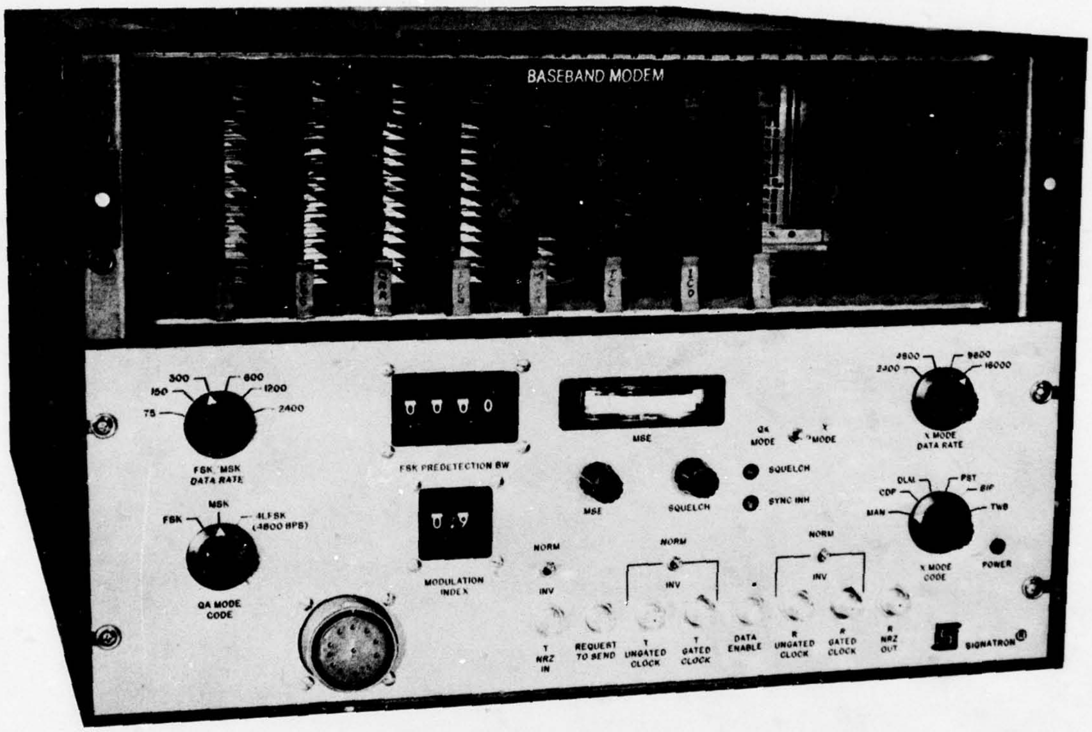


VHF Data Modem/Baseband
Data Signal Converter

SECTION 1
INTRODUCTION AND SUMMARY

1.1 Background of the Program

Tactical VHF radios, while originally used only for FM voice transmission, will be used in the future for the transmission of digital data. In order to evaluate the performance of transceivers with candidate modulation techniques, a means of converting binary data into a format which can be utilized by these radios is required. The VHF Data Modem/Baseband Data Signal Converter (or Baseband Signal Converter, for short), performs the conversion for the following modes and binary data rates:

(a) Quasi-Analog Mode: 75, 150, 300, 600, 1200, 2400, 4800 bps.

(b) X-Mode: 2400, 4800, 9600, 16000 bps.

The data and timing interfaces of the Baseband Signal Converter are in accordance with the low level interface criteria of MIL-STD-188C.

1.2 Signal Formats

As shown above, the Baseband Signal Converter has two operating modes, quasi-analog and baseband. In the quasi-analog mode, the data is transmitted through the normal audio channel of the radio, which typically band-limits the data signals to the 500 Hz to 3500 Hz band. In order to transmit data at rates from 75 bps to 4800 bps, the binary data must be converted to a voice band spectrum which is contained with-

in these limits. The quasi-analog mode provides a choice of three signal types.

Selectable Quasi-Analog Mode Signals

(a) Binary Frequency Shift Keying (FSK)
(b) Minimum Shift Keying (MSK)
(c) Four-Level Frequency Shift Keying (4LFSK).

In the X-mode the data is transmitted over a wideband channel, which in general is connected to the transmitter's frequency modulator input through a circuit which typically has a low frequency cutoff in the 25 to 50 Hz range, and a high frequency cutoff in the 20 kHz range. A wide range of X-mode signal formats are provided:

Selectable X-Mode Signals

(a) Manchester
(b) Conditioned Diphase
(c) Pair Selected Ternary
(d) Delay Modulation (Miller Code)
(e) Bipolar
(f) Twinned Binary.

All of the X-mode signals have spectra which either have nulls or minima at DC, thus minimizing the effect of the modulator's low frequency rolloff.

1.3 Baseband Signal Converter Specifications

The specifications of the Baseband Signal Converter are given in Table 1. Two Baseband Signal Converters are required to implement a full duplex data link.

Table I

SPECIFICATIONS

VHF DATA MODEM/BASEBAND DATA SIGNAL CONVERTER

X Mode Data Rates	2400,4800,9600,16Kb/s
X Mode Codes	Manchester, Conditioned Di Phase Pair Selected Ternary Delay Modulation Bipolar Twinned Binary
Quasi Analog Data Rates	
MSK, FSK	75×2^{N}
4LFSK	$75 \times 2^{N+1}$
	$N = 0, 1, 2, 3, 4, 5$
Quasi Analog Modulations	Frequency Shift Keying (FSK) Minimum Shift Keying (MSK) Four Level Frequency Shift Keying (4LFSK)
Digital Inputs	Request to Send (RTS) TX NRZ Data (TNRZ)
Digital Outputs	Data Enable (DEN) Received NRZ Data (RNRZ) Gated Trans.Timing (GTC) Ungated Trans.Timing (TC) Gated Rec.Timing (GRC) Ungated Rec.Timing (RC)
Digital Interface	MIL STD 188C Low Level Interface
Radio Interface	
Push to Talk	Relay Closure Actuated by RTS 150 ms before data

QA Signal Level Output	7 mv into 150Ω
X Mode Signal Level Output	± 1V into 600Ω
FM Predetection Bandwidth Control	
	50 to 3950 Hz in 50 Hz steps
FM Modulation Index	
Selectables	.2 to 1.9 in .1 steps
Front Panel Indicators	
	Squelch (LED)
	Sync Inhibit (LED)
	Mean Squared Error (Meter)
Miscellaneous Controls	
	Squelch Level
	MSE Adjust
	Input Clock Invert
	Output Clock Invert
	Input Data Invert
Power	117 VAC ±10%, 60 Hz, Single Phase, 1.5A Fuse
Operating Ambient	$20^{o}C$ to $30^{o}C$

1.4 Organization of the Report

This report contains four sections. Section 2, Theory of Operation, presents the theoretical basis of data transmission through tactical radios, including spectral analysis, error rate analysis, and code generation algorithms.

The implementation of the Baseband Signal Converter is described in Section 3. Section 3 is an abridged version of the Equipment Description manual, which has been delivered separately. The block diagram of the Baseband Signal Converter is shown, and the partitioning of the system into eight wire wrap cards is described, along with the operation and interconnection of the functional elements which make up the system.

The final section, Section 4, contains the results of tests performed on the two Baseband Signal Converters built and delivered under this contract. Photographs of baseband and RF spectra are shown, and compared to the theoretical spectra. The measured spectra agree with the theoretically predicted spectra. Error rate curves are presented for all modes at a variety of data rates. The error rate curves can be summarized as follows:

1. The measured X-mode error rates are in excellent agreement with the predicted values. There is little data rate dependence, and the implementation loss appears to be negligable.

2. The measured Quasi-Analog mode error rates show an implementation loss of about 2 dB for MSK and binary FSK, with little data rate dependence. The implementation loss of 4LFSK is data rate dependent, being about 2 dB at 600 bps and 7 dB at 2400 bps, when the modulation index is 0.8/3, or 0.27. In

order to operate at 4800 bps it is necessary to decrease the modulation to 0.1, which results in a signal-to-noise ratio degradation of about 15 dB.

3. There is no significant performance difference between the two delivered units.

SECTION 2
THEORY OF OPERATION

2.1 Data Transmission Through Tactical FM Radios

The error rate performance of digital data transmitted through FM radios can be degraded by several factors. First, the baseband, be it either quasi-analog or X-mode, will have phase and amplitude characteristics which may introduce intersymbol interference in the modulator input data and demodulator output data. To analyze these effects it is necessary to know the band limiting characteristics of the baseband and the spectra of the quasi-analog and X-mode signals. Secondly, filtering between the transmitter modulator output and the receiver discriminator input can be the source of nonlinear phase distortion, which modifies the discriminator output. Thirdly, multipath, which can be either fixed or time-varying, will both cause fading and intersymbol interference. The fading can reduce the received signal level, while the intersymbol interference will cause nonlinear distortion in the discriminator output. Both of these effects can increase the error rate and degrade the performance of the timing circuitry. The final factor to be considered is additive noise, which can be Gaussian receiver noise or impulsive atmospheric noise.

Military FM radios do not operate in an isolated environment. Instead, they are parts of networks, which have specific frequency allocations. In order to minimize the RFI problem, frequencies are allocated which have a low probability of causing in-band and adjacent channel interference. For purposes of standardization, the VHF FM band is divided into 25 kHz wide channels. New radios will use the 25 kHz channel for

both analog speech and 16 kbps data. Older radios use 50 kHz channelization for speech and data. The shape of the transmitted spectrum is therefore an important consideration in signal design.

2.1.1 Typical FM Radio Characteristics

Figure 1 shows the block diagram of a "typical" tactical FM radio. In a real radio many circuits would be shared by the transmitter and receiver, but for clarity these are not shown on the figure. In this section we describe the characteristics of typical radios and interfaces, including specific references to three radios, namely the RT-524/VRC, R-442/VRC, and AN/PRC-77.

The radio is divided into two functional parts, namely the transmitter and the receiver. The transmitter has two operating modes, the normal audio mode and the X-mode. The normal audio mode is energized by the operator pushing the push-to-talk switch, PTT, thereby turning on transmitter power, disconnecting the receiver from the antenna and connecting the transmitter to the antenna. The audio input goes to the audio amplifier, where it is filtered to the 500 Hz - 3500 Hz band. The amplifier sums the filtered audio signal with a 150 Hz sidetone, used at the receiver for squelch. The amplifier output then drives the frequency modulator, whose output is mixed with the local oscillator to RF, amplified, and transmitted. In the X-mode the input signal bypasses the audio amplifier and drives the frequency modulator directly. The bandwidth of the X-mode input is nominally 30 Hz to 16 kHz, and the 150 Hz sidetone is not generated. Squelch is not a normal X-mode function.

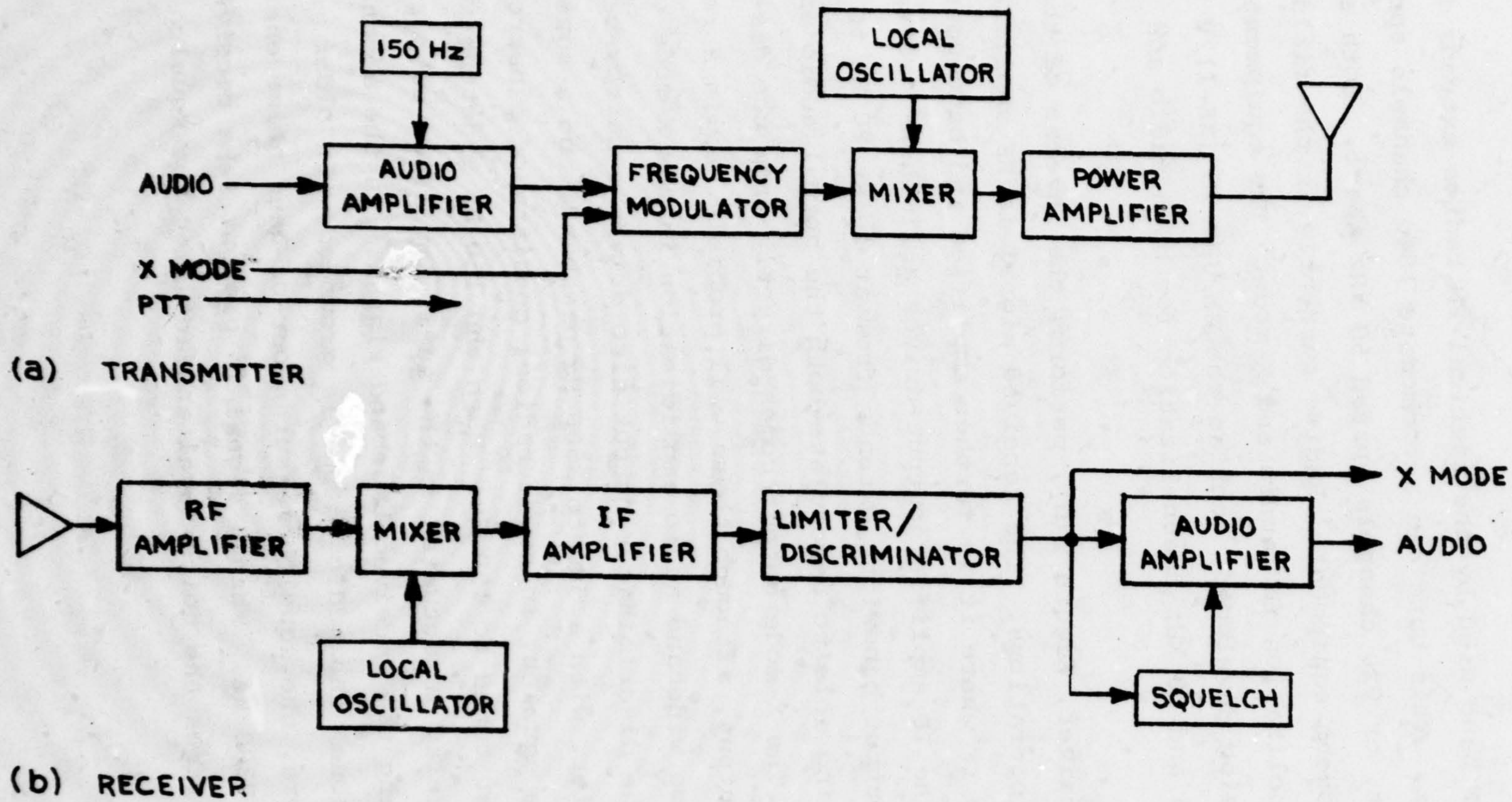


Fig. 1. Typical Tactical FM Radio

The VHF band used by the tactical FM radios extends from 30 to 76 MHz. This band can accommodate 1840 channels spread 25 kHz apart, or 920 channels spread 50 kHz apart. Both are found in present equipment. Radios currently in the field use 50 kHz channelization for audio and X-mode. The equipment currently being developed, such as the AN/URC-78 (XE-1)/V will have both 25 and 50 kHz channelization for both audio and X-mode.

The receiver, Figure 1(b), performs the inverse of the transmitter operations. The received signal is RF amplified and mixed to IF where it is further amplified and band-pass filtered. The IF amplifier output drives a limiter-discriminator, whose output bandwidth extends from at least 20 Hz to 16 kHz, thus being able to demodulate both the normal audio and the X-mode. The X-mode output comes directly from the discriminator output, although there will probably be, in a real receiver, some wideband audio amplification in the X-mode output line. The discriminator output also drives a narrow-band audio amplifier, with a 500 to 3500 Hz passband, and a squelch circuit. The squelch, in newer radios, consists of a narrow band amplifier tuned to 150 Hz. When the 150 Hz sidetone is detected, a relay is actuated in the audio amplifier output, thereby passing the received baseband signal. In the absence of a 150 Hz tone, the audio output is grounded. In older radio receivers, the squelch measures out-of-band noise energy, typically at 7300 Hz. When a signal is present, the receiver quieting decreases the noise level and actuates the squelch relay.

2.1.2 Data and Timing Interface Characteristics

The data and timing interfaces are in accordance with standard low-level interface criteria of MIL-STD-118C, summarized as follows:

2.1.2.1 Transmitter

The transmitter open circuit voltage is positive and negative 6 ±1 volts, with 0.5% ripple and less than 10% unbalance. The transmitter source impedance is less than 100 ohms, with short circuit current less than 0.1 ampere. The rise and fall times are to be between 5 and 15% of the unit interval. For the case of multiple rates, the required waveshaping shall be accomplished at least at the highest modulation rate.

2.1.2.2 Receiver

The minimum input resistance of a single receiver device shall be 6000 ohms. The input capacitance shall not exceed 2500 pf. The receiver sensitivity is such that the maximum operating current required (that current at which the device changes its state from mark to space or vice versa) shall be 0.001 amperes with a minimum input resistance of 6000 ohms.

2.1.2.3 Clock

Clock timing shall be delivered at twice the data modulation rate. The stability of synchronizing or clock timing shall be sufficient to insure that synchronism is maintained within ±25% of the unit interval between transmitted and received signals for periods of not less than 100,000 consecutive

seconds (1 part in e x 10^5 x mod rate). Accuracy of the clock timing shall be such as to assure that the actual modulation rate is within 1 part in 10^7 of the nominal rate.

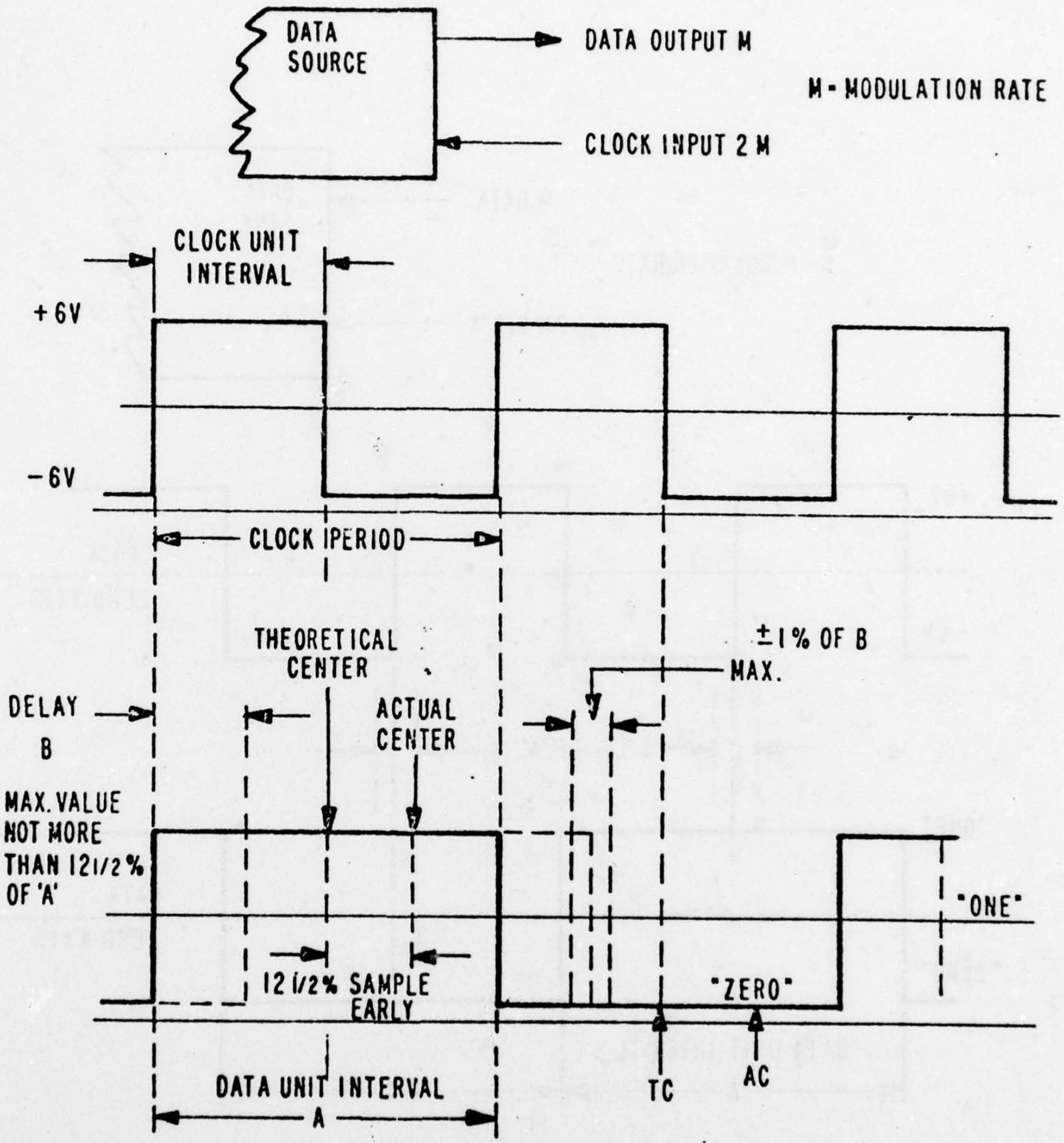
All data transitions emitted by a source under direct control of external clock shall occur on (be caused by) negative to positive transitions of that clock. The design objective is a minimum delay between the clock transition and the resulting data transition, but in no case shall this delay exceed 12.5% of the duration of the data unit interval. For each equipment, once this delay is fixed in hardware, it shall be consistent within ±1% of itself for each clock transition which causes a data transition. These delay limits shall apply directly at the source interface. (See Fig. 2.)

Sampling of the data signal by the external clock at a sink interface shall occur on (be caused by) positive to negative clock transitions. (See Figs. 2 and 3).

When the clock is used for controlling intermittent data transmission, data may not change state except when requested by a negative to positive clock transition. The quiescent state of the clock shall be at negative voltage. The quiescent state of the data shall be that state resulting from the last negative to positive clock transition. (See Fig. 4.)

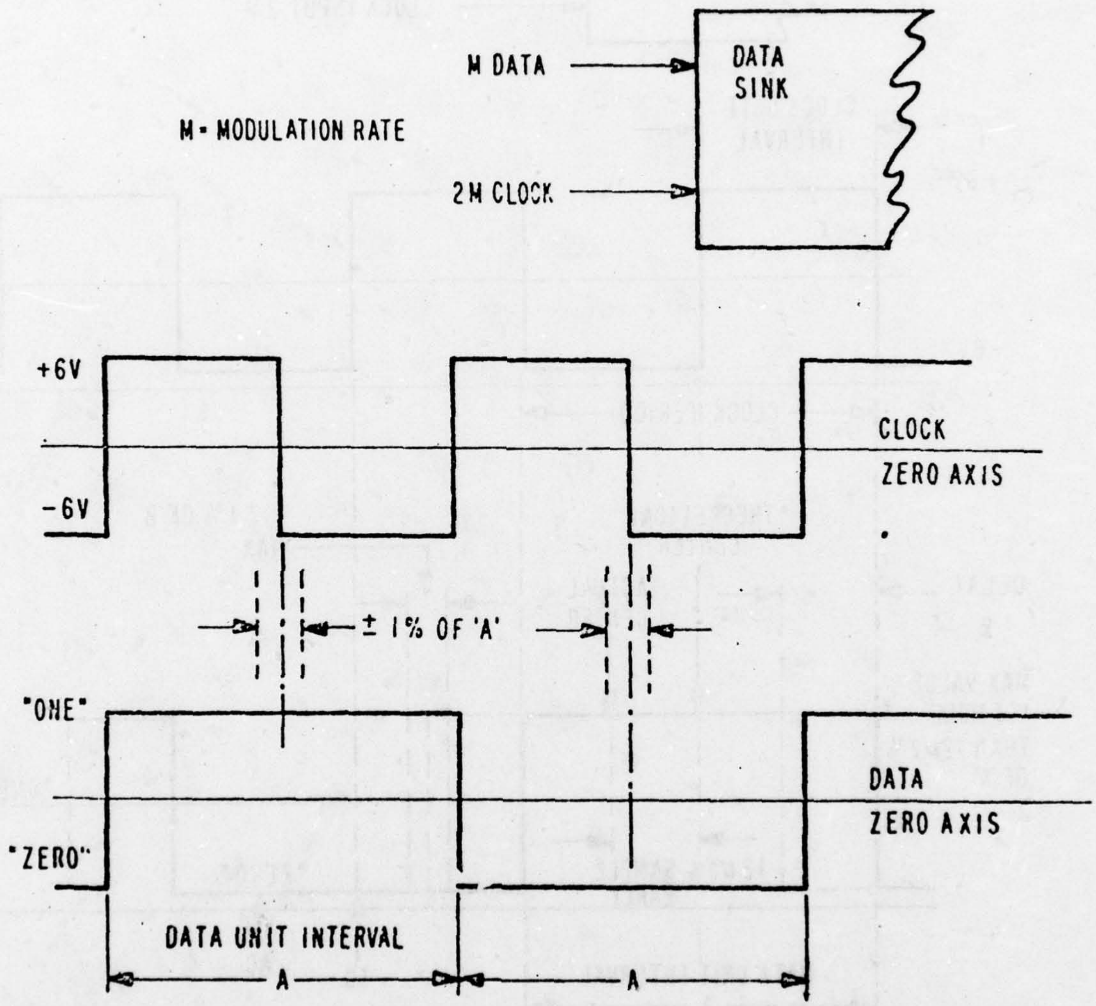
2.1.3 <u>Specific Radio Characteristics</u>

The VHF Data Modem/Baseband Signal Converter is to work with the Radio Set RT-524/VRC, Radio Receiver R-442/VRC, and Radio Set AN/PRC-77. In this section we will discuss the characteristics of these radios which are relevant to the data modem.



The source data signal transition shall occur consistently within 12½% of the duration of the data unit interval from the clock input transition from negative to positive. Whatever phase delay is chosen, the equipment design shall insure that data transitions are released with a consistent phase delay not to exceed plus or minus one percentrof that delay. Measurements shall be made and these limits apply directly at the source interface.

Fig. 2. External Directly Related Source Clock/Data Relationship (From MIL-STD-188c)



The input data signal to the sink shall be sampled consistently within ± 1% of the duration of the unit interval of the data signal assuming the modulation rate of the clock is twice that of the data. Positive to negative clock transitions shall be utilized for sampling of the data.

Fig. 3. External Directly Related Sink Clock/Data Relationship (From MIL-STD-188C)

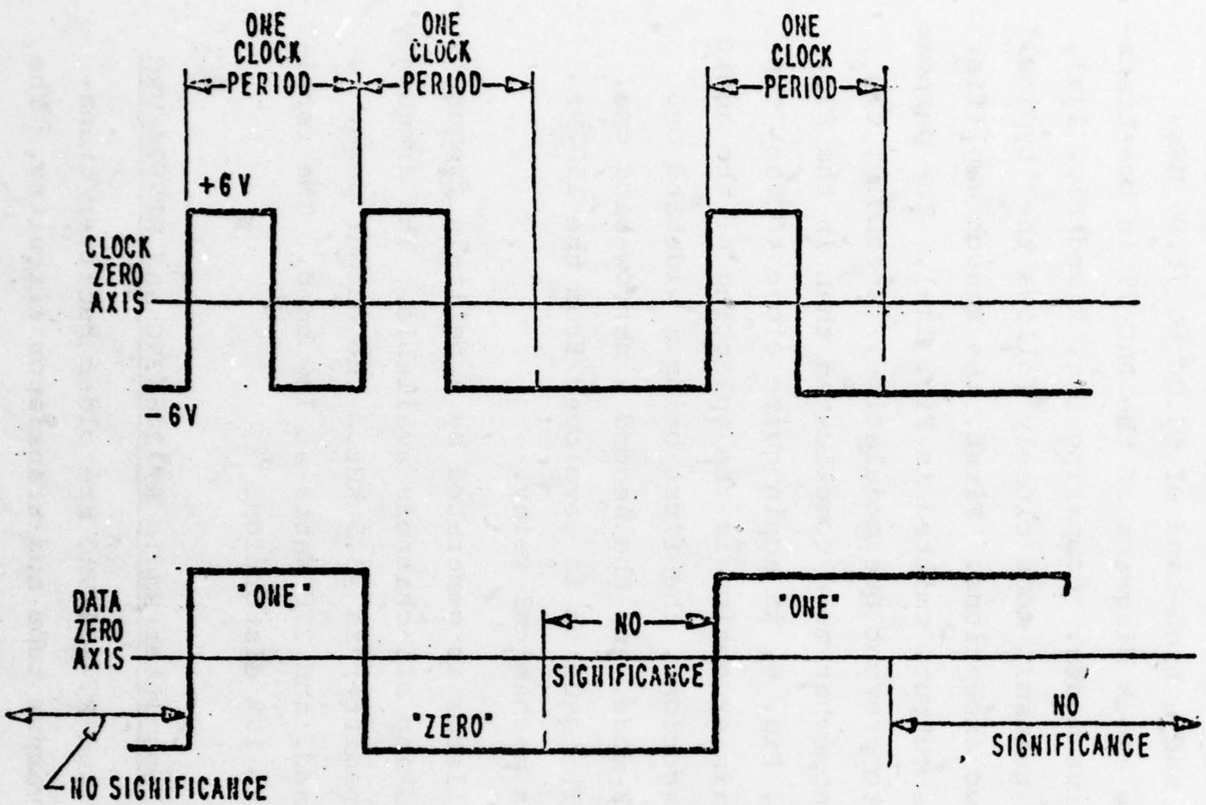


Fig. 4. Standard Interface Clock/Data Phase Relationship
(From MIL-STD-188c)

2.1.3.1 Radio Set RT-841/PRC-77

The PRC-77 is an all solid-state receiver/transmitter operating in the 30-75.95 MHz band. It contains 920 channels spaced at 50 kHz intervals, and divided into a low-band of 30.00 to 52.95 MHz, and a high-band of 53.00 to 75.95 MHz. Figures 5 and 6 show block diagrams of the PRC-77 in the transmit and in the receive modes. Comparing Fig. 5 and Fig. 1(a), it is seen that the transmit mode closely follows the "typical" transmitter, with two exceptions. First, the speech amplifier has a limiter at its output, omitted in Fig. 1(a). The purpose of this limiter is to prevent overmodulation. Secondly, the mixer circuitry is somewhat more complicated than in the Fig. 1(a). The receiver, Fig. 6, is again quite close to the "typical". Here, the main exception is the splitting of the audio amplifier into two sections, the first being a wideband one for both audio and X-mode, and the second a narrow-band one, simply for audio. The squelch is developed from the 150 Hz tone, and operates a mechanical relay.

The local oscillator is generated by a built-in frequency synthesizer, thus making all channels available. The frequency accuracy of the transmitter is ±3.5 kHz. The output power is 1.5 Watts at high band, and 2.0 Watts at low band. The receiver is rated at 5% to 10% distortion.

2.1.3.2 Receiver-Transmitter Radio RT-246/VRC and RT-524/VRC

The RT-246/VRC and RT-524/VRC are older receiver/transmitters, utilizing vacuum tube and transistor circuitry. The transmitter has a 35 Watt high-power output, and 4 Watt low-power output. The block diagram of the radio is similar to a "typical" set. The transmitted frequency is controlled by

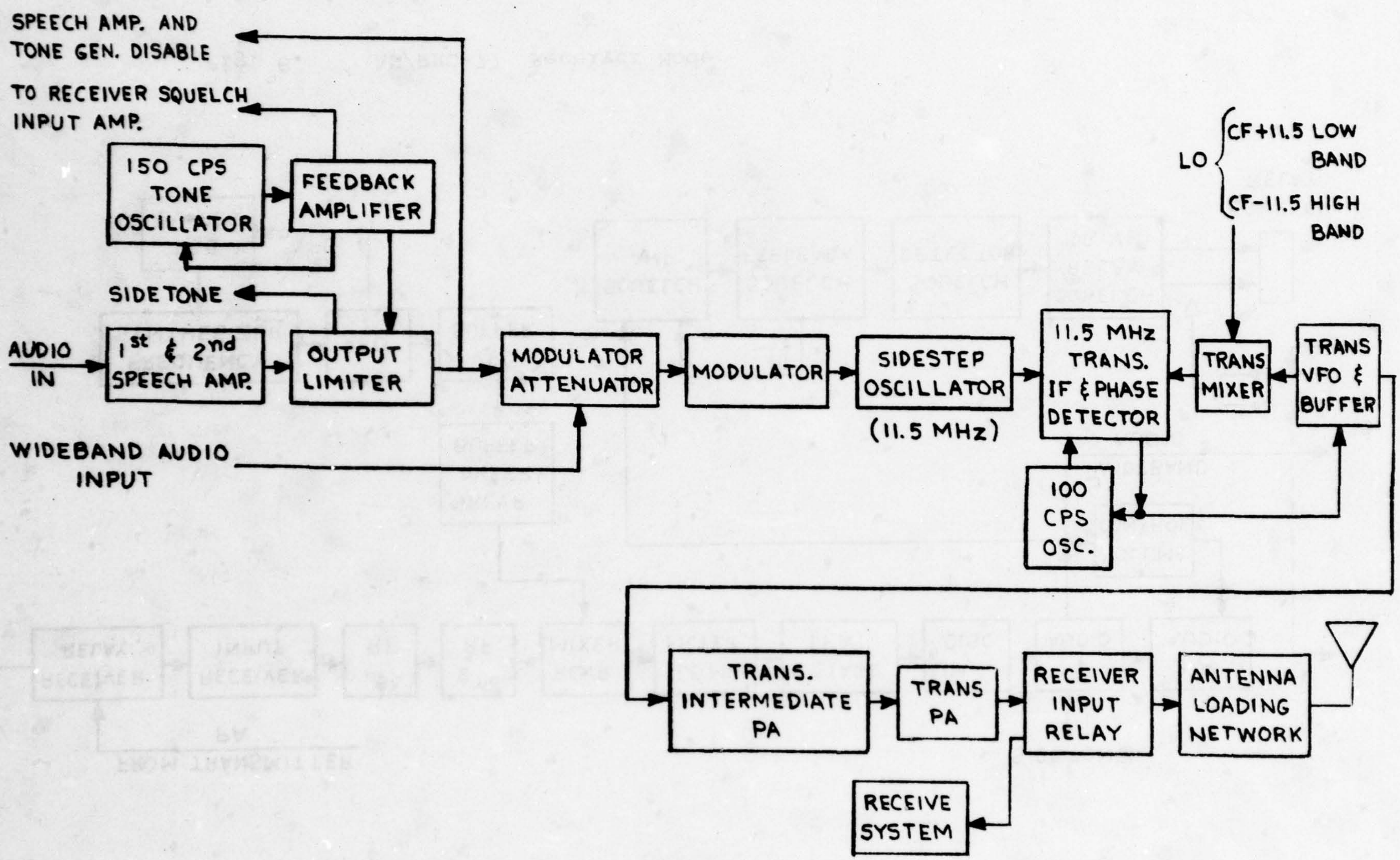


Fig. 5. AN/PRC-77 Transmit Mode

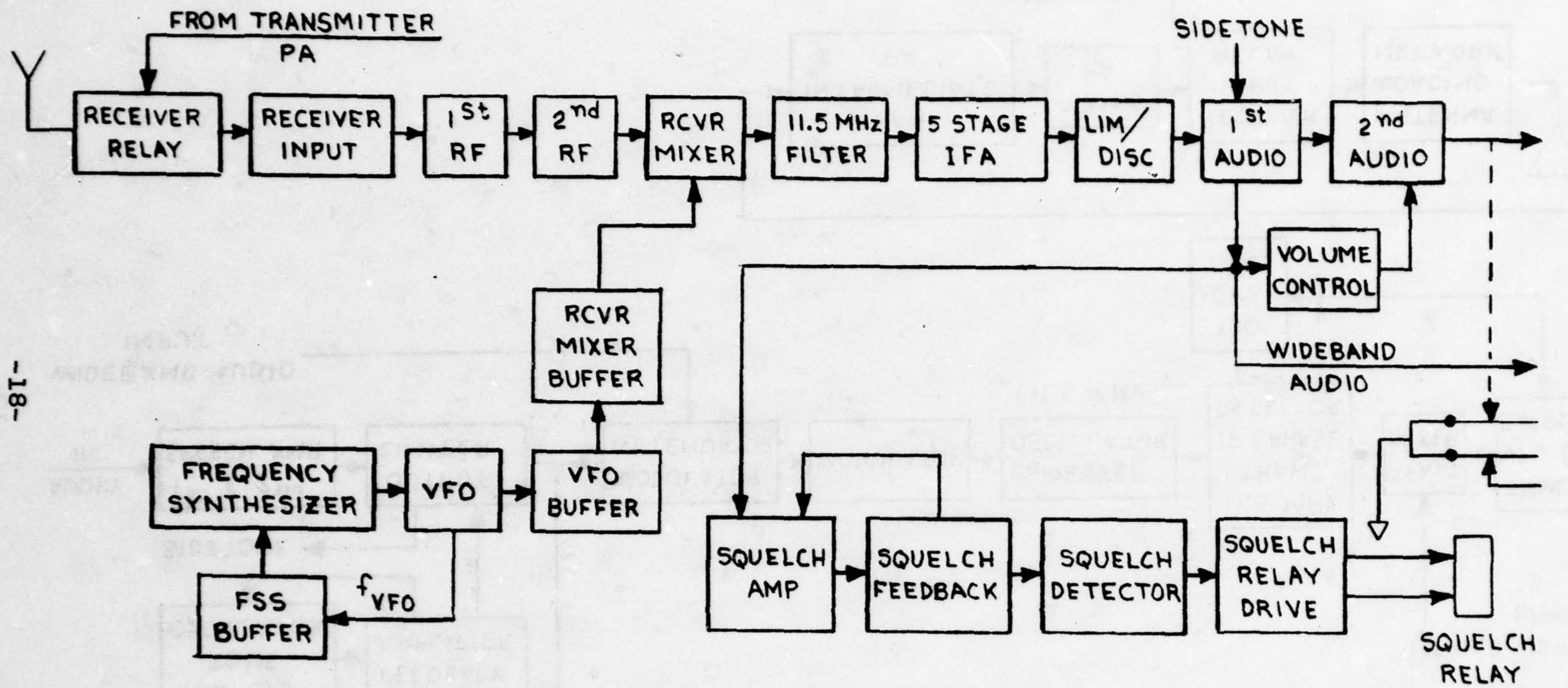


Fig. 6. AN/PRC-77 Receiver Mode

Hartly oscillator, with ganged pretuned LC tuning and crystal discriminator AFC. The transmitter generates a 150 Hz squelch sidetone.

The RT-246 and RT-524 both use the R-422/VRC receiver.

2.1.3.3 Radio Receiver R-422/VRC

The radio receiver R-422/VRC is tunable over the 30.00 to 75.95 MHz range, in either 50 or 100 kHz increments. Like the RT-246/VRC, the local oscillator is a tuned LC oscillator, stabilized by a crystal reference AFC system. Frequency selectivity is achieved by the use of two plug-in crystal filters, one after the mixer and the other after the second stage of a five stage IF amplifier. Different filters are used for normal and for wideband operation -- for normal operation the receiver has a 32 kHz 6 dB bandwidth, while for wideband it has an 80 kHz 6 dB bandwidth. The discriminator is a Travis type, utilizing two tuned circuits. The audio amplifier has a 300 to 3000 Hz passband. Therefore in the X-mode the output must come from either the discriminator output directly, or from the earlier stages of the audio amplifier, before band limiting is applied to the signal. The squelch can be controlled from the front panel to monitor either a 150 Hz tone ("new") or a 7300 Hz noise band ("old"). The R-422/VRC is rated at 8% distortion.

Tuning of the RT-246/VRC, RT-524/VRC, and R-422/VRC is, as has been noted, by means of pretuned LC circuits instead of through the use of a frequency synthesizer as in newer radios such as the AN/PRC-77. Thus these radios, while potentially able to use all channels, are limited to a small number unless they are retuned.

2.2 Quasi-Analog Techniques

A quasi-analog signal is a digital signal, after conversion to a form suitable for transmission over a specified analog channel (MIL STD 188C, Par. 2.575). In the case of tactical FM radios, the specified channel is the audio channel, with a nominal bandpass extending from 500 Hz to 3500 Hz. Input data, at bit rates of 75 x 2^m baud, where m ranges from 0 to 6, i.e., data rates from 75 to 4800 bps, are to be transmitted over this channel.

The available baseband has a 3000 Hz width. The specification requires that the quasi-analog techniques to be considered include FSK, PSK, DPSK, and MSK. In this section we will discuss quasi-analog transmission in general, and then relate it to these specific modulations.

2.2.1 Quasi-Analog Transmission

Figure 7 shows the essentials of an FM transmitter/receiver voice channel. The audio input signal is bandlimited by a high pass filter/low pass filter sequence, often pre-emphasized, and inputted to the frequency modulator. The FM output is then mixed to RF, amplified, and transmitted. At the receiver the inverse operations are performed, with the input baseband modulation being recovered by the discriminator. The squelch circuit disables the audio amplifier in the absence of an input signal.

The concatenated transmitter/receiver is, from input to output, a baseband channel which has amplitude and phase distortion primarily due to the baseband and IF filters, and distortion in the order of 5% due to implementation nonlinearities.

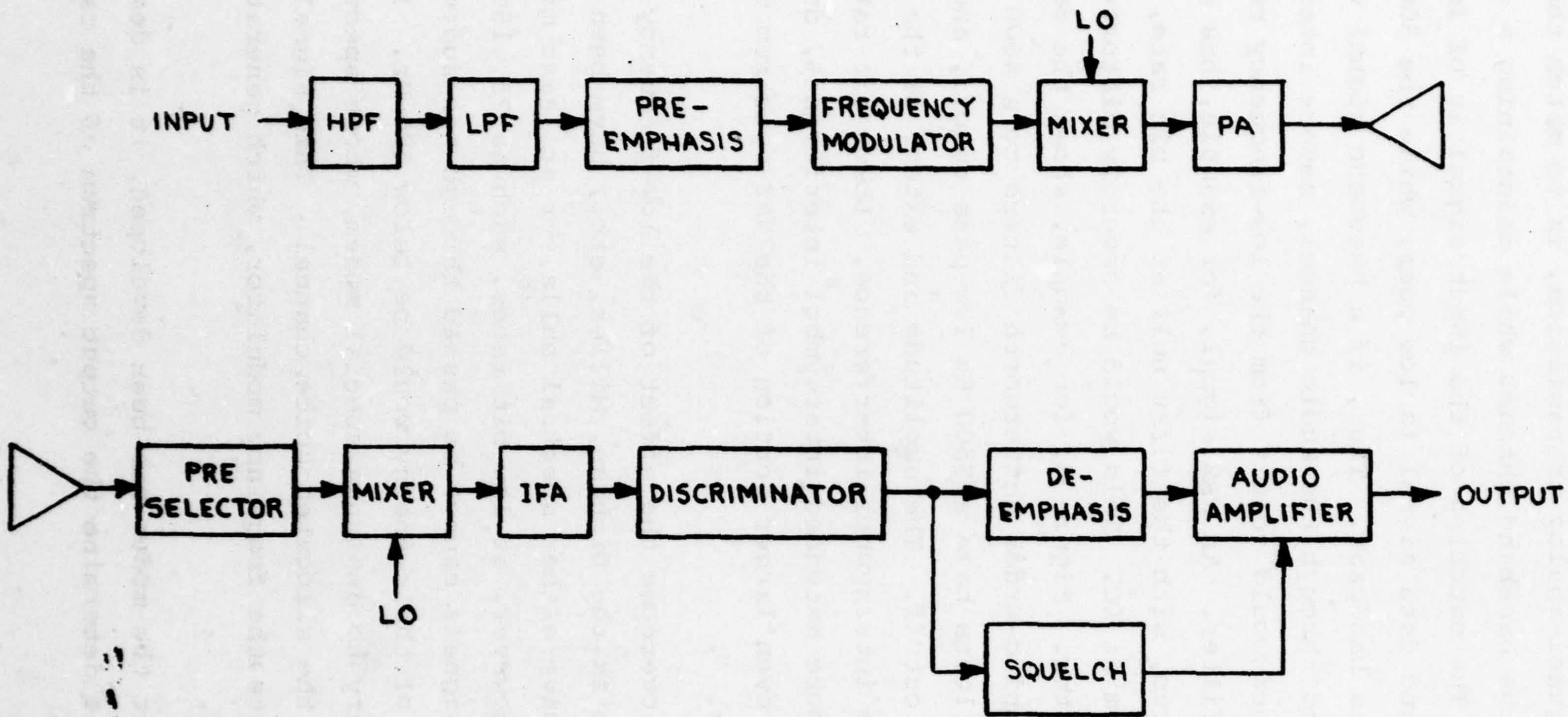


Fig. 7. FM Transmitter/Receiver

The problem, in quasi-analog transmission, is to match the input signal to the baseband channel while maintaining a useful bit error rate. The matching of the input signal is of importance as a baseband data signal is low pass, while the 500 to 3500 Hz channel is bandpass. Thus, if a baseband signal were simply transmitted through the audio channel, severe intersymbol interference would result from the low-frequency cutoff of the highpass filter. An NRZ signal, for example, has a $(\sin x/x)^2$ spectrum, with the first null at the bit rate, and a spectral maximum at DC. This would be severely distorted by the high-pass filter. Figure 8, for example, shows the output of concatenated third-order Butterworth filters to a 4800 bps NRZ input; the filters have a 3500 Hz low-pass cutoff, and a 500 Hz high-pass cutoff. The amplitude and extent of the tail will cause severe intersymbol interference. Lower bit rates would have even more extended intersymbol interference, due to the fact that an even larger portion of the NRZ spectrum would fall below 500 Hz.

In order to overcome the effect of the low-frequency cutoff, signals such as the Diphase, Miller, etc., have been developed which have either spectral nulls, or at least minimum energy, at DC. However, at low bit rates, such as 75, 150, 300 bps, even these signals cannot be passed through the audio channel, as most of their energy would be below 500 Hz. It is therefore necessary to design a special modem, whose spectrum fits well within the allocated voice channel. This signal then is used to modulate the frequency modulator, which generates the output signal.

Assuming that the modem has been developed, it is desirable to be able to determine the output spectrum of the radio.

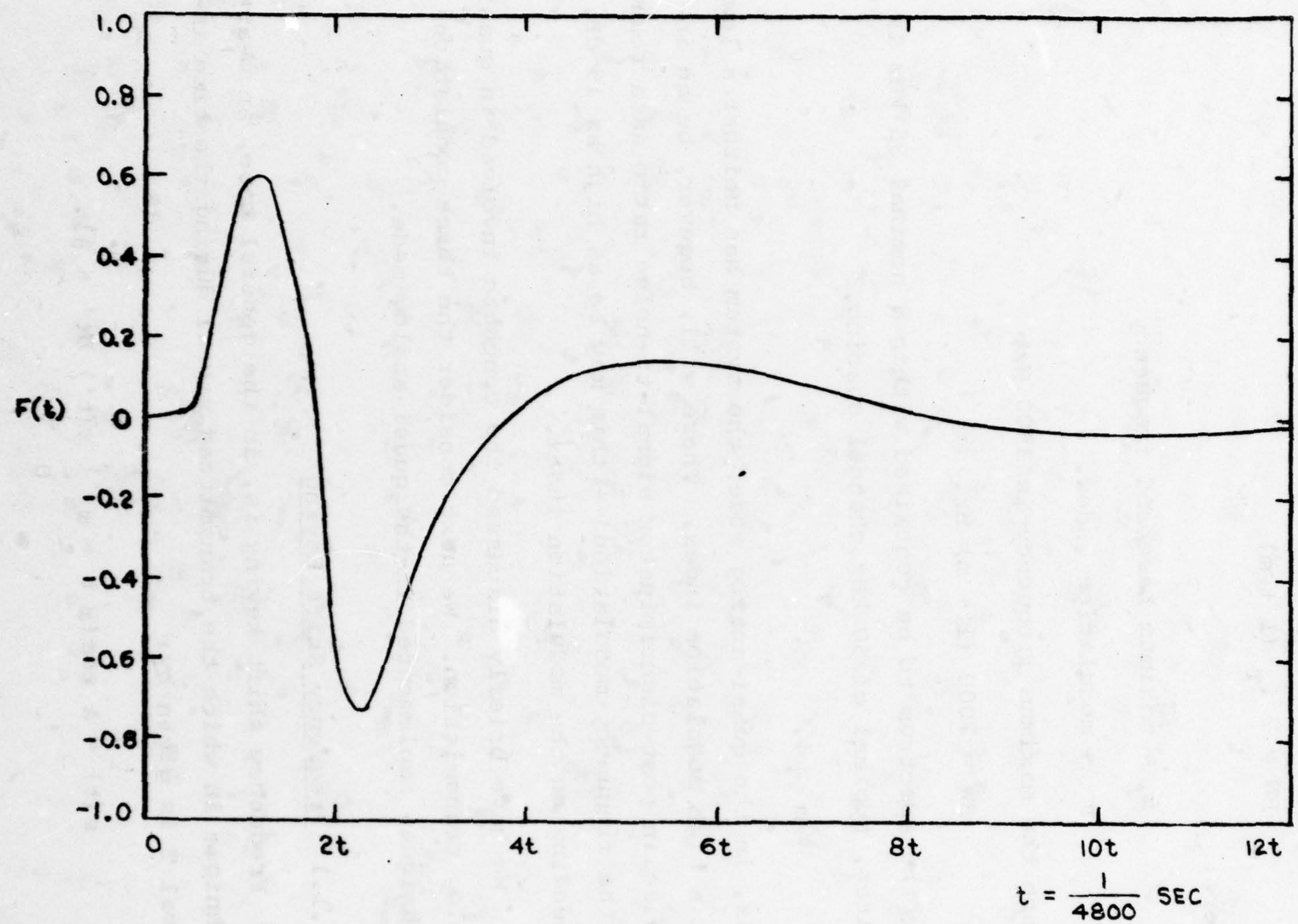


Fig. 8. Pulse Response 4800 bps NRZ Pulse, Quasi-analog Channel

For ease of analysis we will assume that the bandwidth of the transmitted signal can be approximated by Carson's rule:

$$BW \cong 2\ f_T\ (1 + m)$$

where:

f_T = maximum baseband frequency

m = modulation index.

Since the maximum frequency is 3500 Hz,

$$BW \cong 7000\ (1 + m)\ \text{Hz}.$$

For the spectrum to be contained within a nominal 35 kHz IF filter, typical of 50 kHz channel spacing,

$$m \sim 4.$$

Thus, in the quasi-analog mode, the system has neither a low nor a high modulation index. There will, however, be an increase in post-discriminator signal-to-noise ratio as a result of the frequency modulation -- this may be as high as 15 dB, depending on the modulation index.

We have briefly discussed the concepts involved in quasi-analog transmission. We next consider the three modulation techniques implemented in the quasi-analog mode.

2.2.1.1 Frequency Shift Keying

Frequency shift keying is, in the general case, an m-ary technique in which the transmitted signal during the time interval T is given by:

$$s(t) = A\cos\left(\omega_c t + \omega_d \int_0^t x(t')\,dt' + \theta\right).$$

Thus, ω_c is the carrier frequency, ω_d is a scale factor, and θ is an arbitrary phase angle. If x(t) has the representation

$$x(t) = \sum_n a_n \, g(t-nT),$$

that is, a series of pulses of basic waveform g(t) multiplied by data symbols a_n, then

$$s(t) = A \cos(\omega_c t + \omega_d \sum_n \int_0^t a_n g(t-nT)\, dt + \theta).$$

If g(t) is a rectangular pulse of unit amplitude and duration T, then the frequency of s(t) changes every T seconds. Two cases can be considered, one in which the phase is discontinuous, and the other in which the phase is continuous. Phase discontinuous FSK is made by selecting one of M (for M'ary) free running oscillators. Phase continuous FSK is made either by modulating a voltage controlled oscillator (VCO) by the data signal, or by special signal design. Phase continuous FSK is also known as digital FM

In binary digital FM, or binary FM for short, the transmitted frequencies reduce to a mark frequency, ω_m, and a space frequency, ω_s. The magnitude of the difference between these frequencies and the center frequency of the VCO, ω_c, is the frequency deviation, ω_d.

$$\begin{aligned} \omega_d &= |\omega_m - \omega_c| \\ &= |\omega_s - \omega_c| \\ &= |\omega_m - \omega_s|/2. \end{aligned}$$

In the M-ary case, the frequencies are spaced at $2\omega_d$ intervals, located at $\omega_c \pm (2m-1)\omega_d$, m=1,2,...M/2, with the outermost frequencies being $\omega_c \pm (M-1)\omega_d$. A frequency deviation parameter, k, can be defined as

$$k = \frac{\omega_d T}{\pi} ,$$

where T is the duration of the M-ary baud. Since $\omega_d = 2\pi f_d$, $f_d = k/2T$. Therefore the mark-space frequency separation is

$$|f_m - f_s| = \frac{k}{T} .$$

If k is kept constant, the bandwidth of an M-ary FM signal increases approximately proportionally with M. If k decreases as 1/M, the bandwidth can be kept approximately constant as M increases.

Figure 9 shows the spectrum of binary NRZ FM for three values of k, namely 0.5, 0.6 and 0.7. The Y axis is in dB and the X axis the normalized parameter $(\omega-\omega_c)/2\pi T$, where ω is radian frequency, ω_c is the carrier frequency and T is the bit duration in seconds. It is seen that the first lobe of the spectrum is contained in a frequency of the order of 0.6 to 0.75 the bit rate, thereby occupying a quasi-analog bandwidth of 1.2 to 1.5 times the bit rate. As the available spectrum is 3 kHz, binary FM is limited to approximately 2 kbps to 2.4 kbps in the quasi-analog mode.

The bandwidth required to transmit 4800 bps using binary FSK would be 5300 Hz. Therefore, in order to transmit 4800 bps, at least four-level modulation must be used. With four frequencies, two bits are encoded into one symbol, so that

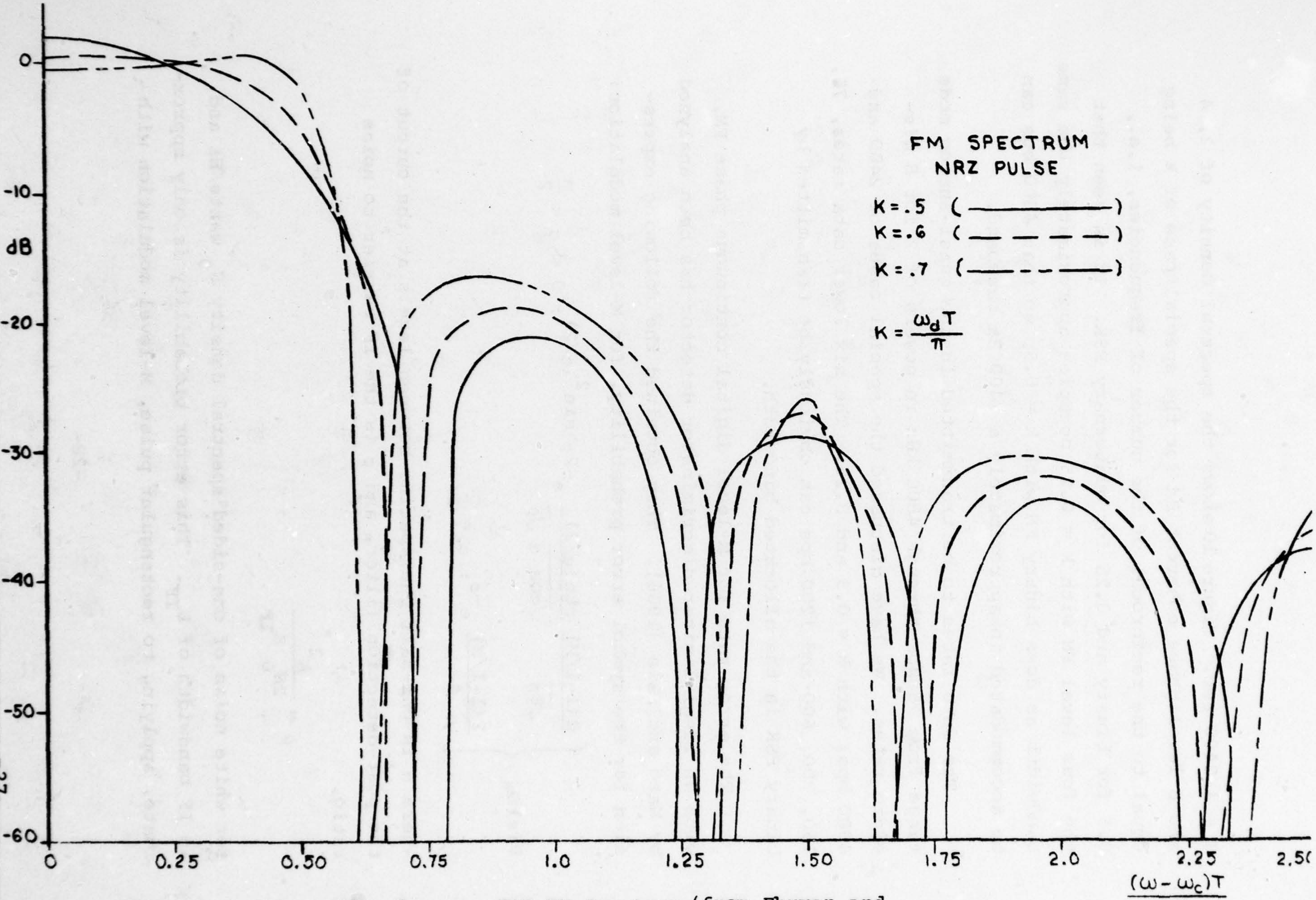


Fig. 9. NRZ FM Spectrum (from Ehrman and Westerberg, 1976)

$T = 1/2400$ sec. Figure 10 shows the spectral density of 2, 4 and 8 level phase coherent FM for the special case of k being equal to the reciprocal of the number of frequencies, i.e., 0.5 for binary and 0.25 for quaternary FSK. It is seen that the four level FM with $k = 0.25$ occupies approximately the same bandwidth as does binary FM with $k = 0.5$, so that 4800 bps can be accommodated in approximately a 2400 Ha baseband.

The data rates to be transmitted in the quasi-analog mode range from 75 bps through 4800 kHz in powers of 2 for 8 discrete rates. We have discussed the special cases of 2400 and 4800 bps, with $k = 0.5$ and 0.25. The six lower data rates, 75, 150, 300, 600 and 1200 bps can obviously be transmitted by binary FSK in the allocated bandwidth.

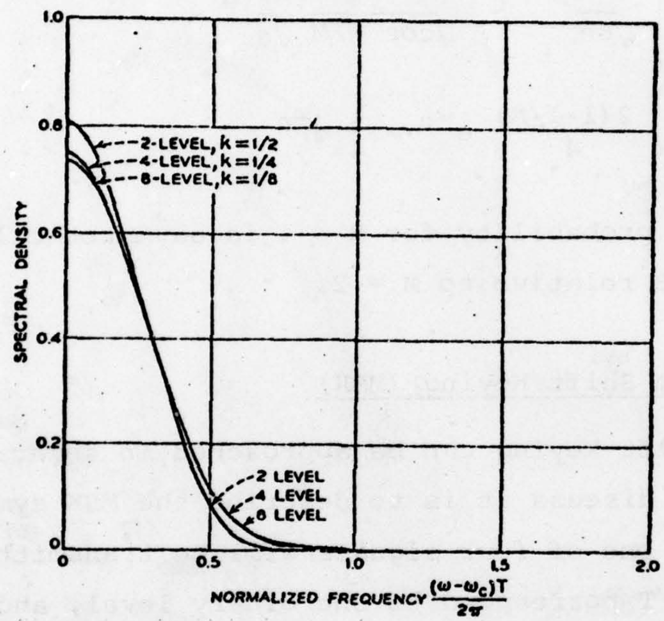
The error rate for M-level digital continuous phase FM, with an ideal limiter-discriminator detector has been analyzed by Mazo and Salz (1966). They obtained the following expression for the symbol error probability, for M-level modulation:

$$P(\epsilon) \approx \begin{cases} \dfrac{2(1-1/M)}{\sqrt{8\pi}} \, \dfrac{\operatorname{ctn}(\alpha/2)}{\sqrt{\cos\alpha}\,\sqrt{\rho}} \, e^{-2\rho \sin^2(\alpha/2)}; & 0 < \alpha < \dfrac{\pi}{2} \\ \dfrac{2(1-1/M)}{4} \, e^{-\rho}; & \alpha = \dfrac{\pi}{2} \end{cases}$$

where α is one-half the spacing between levels at the output of the post-detection filter, and ρ is the IF carrier to noise ratio,

$$\rho = \frac{A^2}{2N_0 \, B_{IF}}$$

for white noise of one-sided spectral density N_0 watts/Hz and an IF bandwidth of B_{IF}. This error probability is only approximate, applying to rectangular pulse, M-level modulation with



$$k = \frac{\omega_d T}{\pi}$$

Fig. 10. Spectral density for k = 1/No. of levels. (from Lucky, Salz, Weldon, 1968)

integrate and dump post-detection filtering and ignoring all intersymbol interference effects. When the modulation index k is set to 1/M, the frequency shifts are $\pm\frac{\pi}{MT}$, $\pm 3\frac{\pi}{MT}$, etc. After T seconds integration, one half the level spacing is $\alpha = \pi/M$. Substituting this yields

$$P_e(\mathrm{NRZ}) = \begin{cases} \dfrac{2(1-1/M)}{\sqrt{8\pi}} \; \dfrac{\mathrm{ctn}(\pi/2M)}{\sqrt{\cos \pi/M}\,\sqrt{\rho}} \; e^{-2\rho \sin^2(\pi/2M)} & M > 2 \\[2ex] \dfrac{2(1-1/M)}{4}\, e^{-\rho} = \dfrac{1}{4}\, e^{-\rho} & M = 2. \end{cases}$$

Thus, the error probability for M = 4 is asymptotically degraded by 4.5 dB relative to M = 2.

2.2.1.2 Minimum Shift Keying (MSK)

Minimum shift keying can be approached in several ways. A simple way to discuss it is to describe the MSK symbol. For each input bit, one of four signals will be transmitted. The signals $\pm\sin \pi t/T$ correspond to one binary level, and the signals $\pm\sin 2\pi t/T$ correspond to the other. The $\pm$ sign is chosen so that the transmitted signal is phase continuous.

With this description, it is clear that MSK is simply coherent FSK with a modulation index of 0.5. The difference between MSK and coherent FSK arises in its detection. If coherent phase detection is performed, using two adjacent bauds,

$$P_e = \frac{1}{2}\,\mathrm{erfc}\,\sqrt{\rho}\ .$$

This error rate is the same as ideal binary PSK.

2.3 X-Mode Techniques

In this section we will describe the X-mode signalling techniques in detail and present sample wave shapes as well as the baseband spectra for each of the techniques. We will briefly discuss salient characteristics of the signals and their spectra.

1) Dicode
2) Twinned Binary
3) Bipolar
4) Delay Modulation (Miller Coding)
5) Paired Selected Ternary (PST)
6) Conditioned Diphase
7) Manchester Coding.

2.3.1 Description of Techniques

In the following description, the binary bit sequence which is to be encoded will be represented by a_n which has the value of ± 1. The bit rate is W bits per second, and the duration of a bit is $T (=1/W)$ seconds. Figure 11 shows sample wave shapes for each of the techniques.

2.3.1.1 Dicode

In dicode transmission, a positive pulse is sent whenever there is a minus-to-plus transition of the input sequence; a negative pulse for a plus-to-minus transition; and zero when there is no transition. Dicode can be used for asynchronous data, the only requirement being that signal transitions must be no closer than the duration of the dicode pulse. The dicode spectrum has no energy at DC. The dicode pulse is rectangular, with its width equal to T, the bit duration.

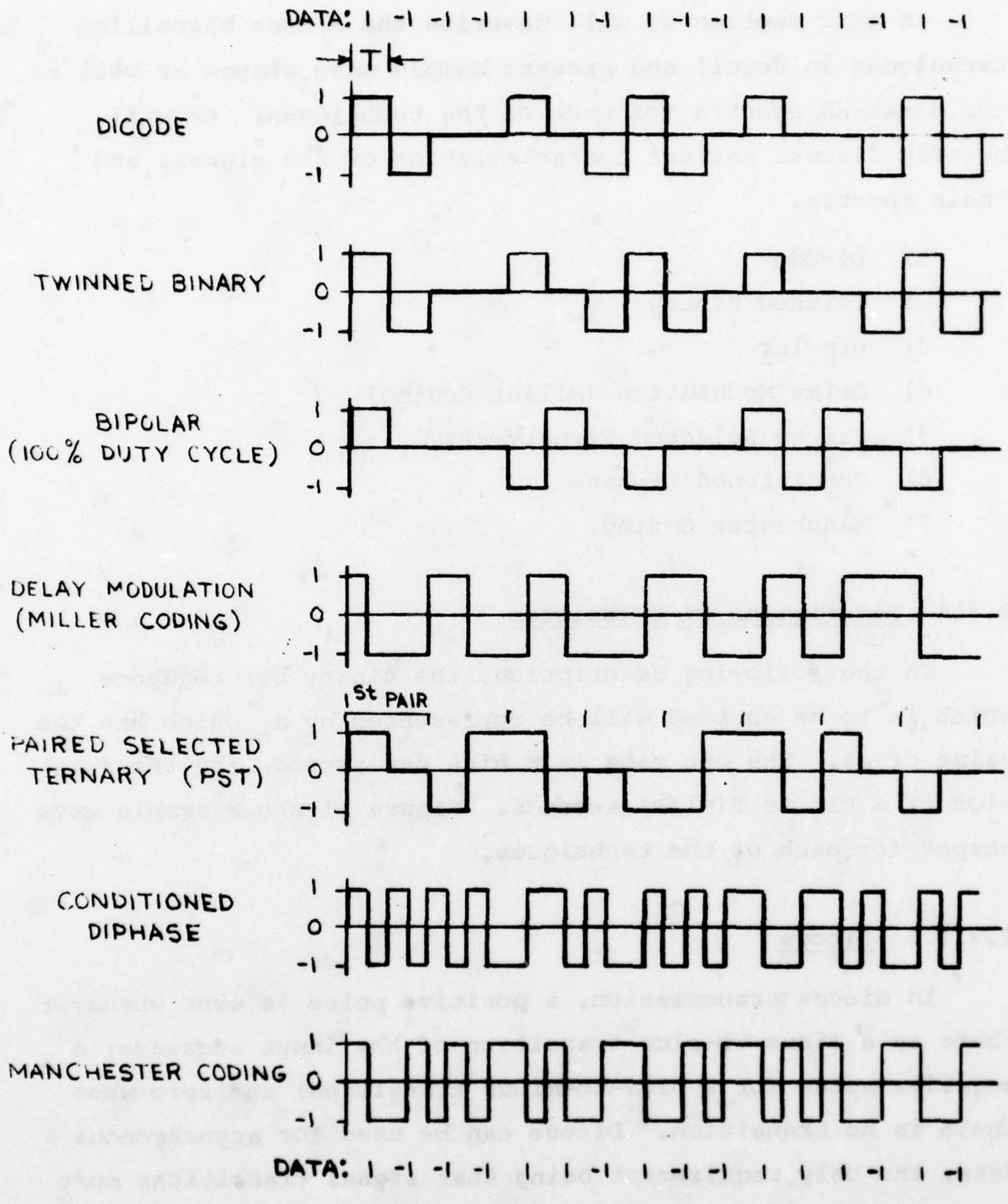


Fig. 11. Sample Baseband Waveshapes

2.3.1.2 Twinned Binary

Twinned binary is obtained by delaying the input bit stream by one bit, subtracting, and normalizing the difference:

$$b_n = (a_n - a_{n-1})/2.$$

The resulting sequence has three values: +1, 0 and -1. Pulses of amplitude +1, 0 and -1 are used to represent these values. The pulse shape is a square pulse of length T.

With this pulse shape, it is readily verified that the twinned binary and dicode techniques are identical, since the same truth table is obtained from the dicode transition rule and from the twinned binary delay-and-subtract formula:

a_{n-1}	a_n	dicode	twinned binary
-1	-1	0	0
-1	1	1	1
1	-1	-1	-1
1	1	0	0

2.3.1.3 Bipolar

Bipolar signals are obtained by transmitting 0 on a -1 input, and a positive or negative pulse on a +1 input, with polarity of the nonzero pulses strictly alternating. The pulse used is conventionally a square pulse with its length between T/2 and T, or duty cycle between 50% and 100%. Because the polarity of the pulses is strictly alternating, the DC component will be zero. Since the polarity of the first pulse is not specified, there is a ±1 ambiguity.

This coding can also be accomplished by precoding, to form the sequence $c_n = c_{n-1} \oplus a_n$. The $\oplus$ symbol represents exclusive-or. The c_n sequence is coded by the rules of twinned binary or dicode, both of which will yield the same sequence as bipolar coding, except possibly for the duty cycle and the ±1 ambiguity. Alternately, twinned binary can be obtained by differential precoding, $d_n = a_n \oplus a_{n-1}$ and applying bipolar coding to the d_n sequence.

2.3.1.4 Delay Modulation or Miller Coding

Delay modulation produces a binary output rather than ternary as the previous techniques produced. Delay modulation is obtained from the following rules:

(1) The first half baud is the same as the previous half baud if the present symbol is +1 or if the previous symbol was a +1 or both. Otherwise, it has the opposite polarity.

(2) The second half baud is the same as the first half if the present symbol a_n is -1; otherwise (if a_n = +1), the second half baud has the opposite polarity from the first half baud.

It will be seen that the output never goes for more than 2T without a transition. However, it can have a DC component since there are data sequences, such as 110110 which cause the signal to have one polarity more than the other. The spectrum thus has a minimum, instead of null, at DC.

2.3.1.5 Paired Selected Ternary (PST)

Paired selected ternary groups the input bits into pairs,

then encodes the pairs as follows:

INPUT		OUTPUT	
a_{2n}	a_{2n+1}	b_{2n}	b_{2n+1}
-1	-1	-1	1
-1	1	0	±1
1	-1	±1	0
1	1	1	-1

The occurrence of either of the symbols with the ± signs is strictly alternating. This forces the sequence to have zero long term d.c. average, since d.c. contributions can arise only from end effects. In decoding, the signal must be broken into pairs correctly. If the pairing is wrong, one of the "impossible" pairs: (0,0), (1,1), or (-1,-1) will occur before long, indicating the error.

2.3.1.6 Conditioned Diphase

Conditioned diphase is related to ordinary diphase or Manchester Code (described below). The input is precoded,

$$b_n = b_{n-1} \oplus a_n$$

and the b_n's determine the signs of a train of diphase pulses. The diphase pulse is positive for the first half baud, negative for the second half, and always has a transition at baud center.

2.3.1.7 Manchester Coding

Manchester Coding, or level diphase, is obtained by use of the input sequence directly to modulate a train of diphase pulses. Since the individual pulses have zero DC component,

the conditioned diphase and Manchester Code signals will both be free of any DC component.

2.3.2. X-Mode Baseband Spectra

The baseband power density spectra of all the X-mode data signals are presented in this section, both graphically and analytically. The formulas are listed in Table 2.

There are two cases where different signaling methods give the same spectra. It was pointed out twinned binary and dicode are identical, so of course, they have the same spectrum. This is also the spectrum of 100% duty cycle bipolar. This occurs because dicode and 100% bipolar are related to each other by a precoding operation, and for a random bit sequence (i.e., bits independent, ±1 equiprobable), precoding does not change the statistics of the bit sequence.

The other case is that of Manchester Coding and conditioned diphase. These two signals are also related to each other by a precoding operation, which again does not change the bit sequence statistics if the bits are independent with ±1 equiprobable, so that this pair of signals will have identical spectra.

The baseband spectra appear in Figures 12 and 13. The figures are plotted on the same coordinates to facilitate direct comparison. The power density is plotted on a decibel scale, since this facilitates comparison of the sidelobes as well as of the main lobe. All are normalized to peak signal levels of ±1 and to a bit length of T=1 second. Note that this is not equivalent to equal power since the ternary, with their zero level, will have lower average power. The difference in total power between the signals is never greater than 3 dB.

TABLE 2
BASEBAND SPECTRA

Baseband Technique	Power Spectral Density, $G(f)$ ($\omega = 2\pi f,\ \theta = \pi f T$)
Dicode Twinned Binary Bipolar (100% duty cycle)	$T\left(\frac{\sin^2(\omega T/2)}{\omega T/2}\right)$
Bipolar (50% duty cycle)	$T\left(\frac{\sin(\omega T/4)\ \sin(\omega T/2)}{\omega T/4}\right)^2$
Miller Coding	$\frac{2T}{(2\theta)^2\ [17+8\cos 8\theta]}\left\{23-2\cos\theta-22\cos 2\theta -12\cos 3\theta+5\cos 4\theta+12\cos 5\theta+2\cos 6\theta -8\cos 7\theta+2\cos 8\theta\right\}$
Paired Selected Ternary	$\frac{T}{8}\left[\frac{\sin(\omega T/2)}{\omega T/2}\right]^2(1-\cos\omega T)\ (7+4\cos\omega T + 2\cos^2\omega T)$
Conditioned Diphase Manchester Coding	$T\left[\frac{\sin^2(\omega T/4)}{\omega T/4}\right]^2$

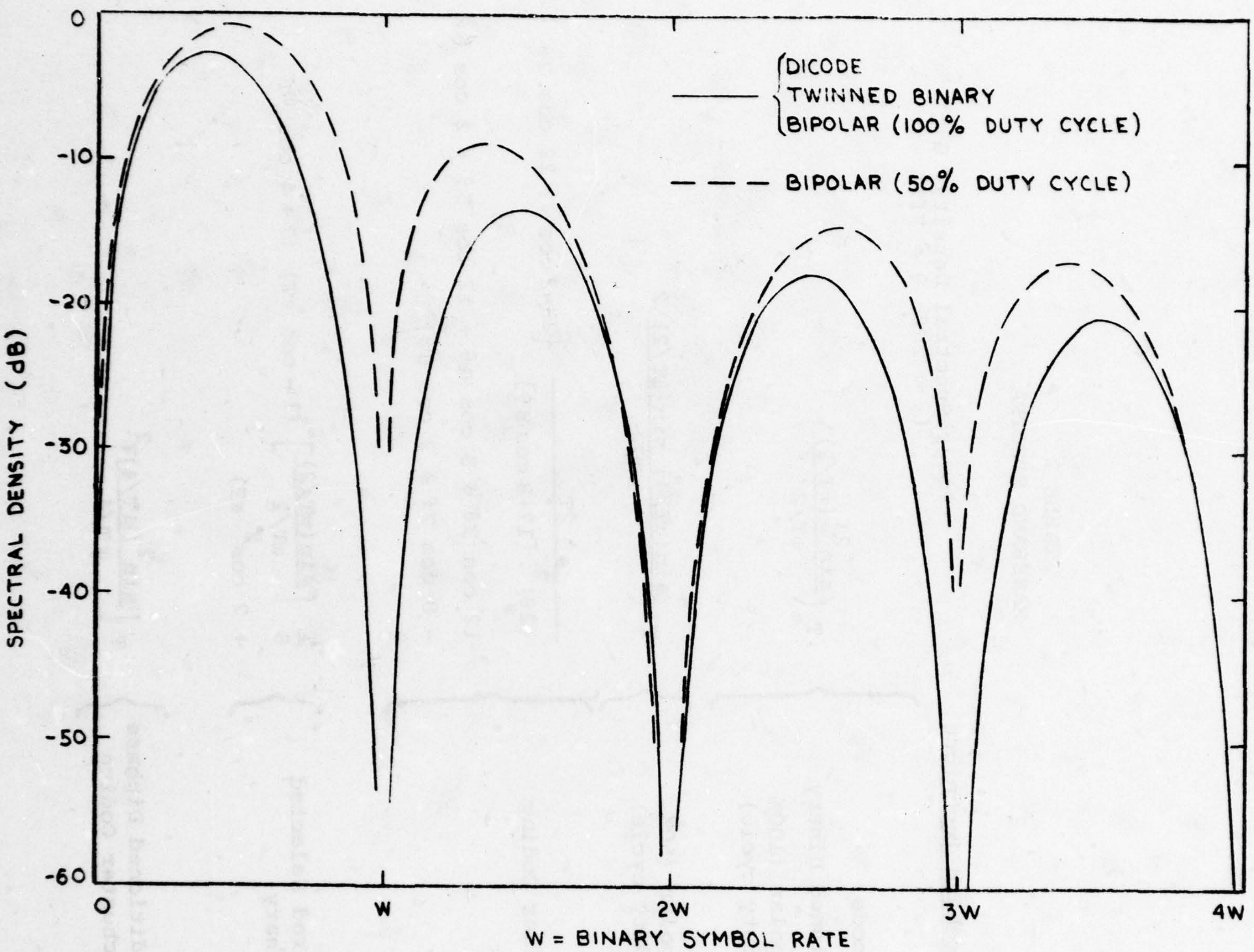


Fig. 12. Baseband Spectra, Dicode, Twinned Binary, and Bipolar

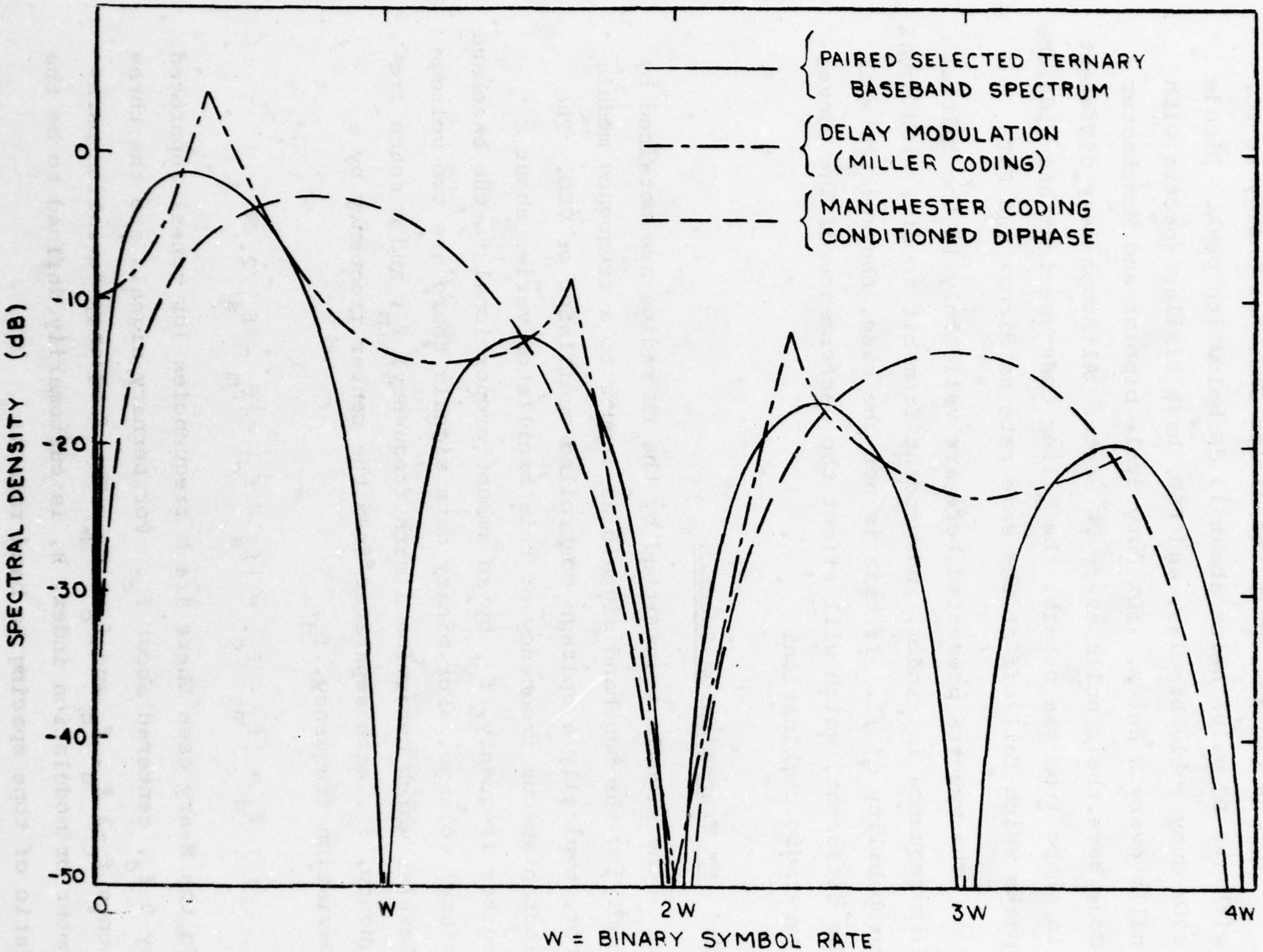


Fig. 13. Baseband Spectra, PST, Delay Modulation, Conditioned Diphase, Manchester Coding

Of the signals, only the Miller Code has energy at DC, with its DC value being about 13 dB below its peak. Dicode 100% duty cycle bipolar, and PST, have similar spectra with nulls every W Hertz. 50% duty cycle bipolar and Manchester Code have their nulls every 2W Hertz. Although very different in shape from the others, the Miller Code spectrum has sidelobe peaks which fall off at the same rate as dicode and PST.

The spectra presented here are valid only if the initial bit sequence is random, independent from bit to bit, with equal probability of ±1. If this is not the case, the spectra will be different, which will affect the performance of the converter-radio combination.

2.4 FM Transmission Spectra

The signals transmitted by the FM radios are obtained by applying the baseband signals as inputs to a frequency modulator, typically a voltage controlled oscillator or VCO. The instantaneous frequency of this oscillator varies about a center frequency, f_c, by an amount proportional to the baseband signal voltage. For binary data signals there are two voltage levels, which determine a mark frequency, f_m, and a space frequency, f_s, each separated from the center frequency by a deviation frequency, f_d,

$$f_d = |f_m - f_c| = |f_s - f_c| = |f_m - f_s|/2.$$

In the M-ary case there are M frequencies (or tones) separated by 2 f_d, centered about f_c. For ternary signals are the three tones f_c-2 f_d, f_c and f_c+2 f_d. The frequency deviation parameter or modulation index, k, is customarily defined to be the ratio of tone spacing to baud rate R:

$$k = \frac{2\ f_d}{R} = 2\ f_d\ T = \frac{\omega_d T}{\pi}\ ,$$

where $\omega_d = 2\pi f_d$.

The FM signal is given by

$$S(t) = A\ \cos\left\{\omega_c t + \int_0^t \omega_d \sum_{n=0}^{\infty} b_n\ g(t'-nT)dt+\varphi\right\}\ ,$$

where

ω_c = $2\pi f_c$ = center frequency (radians)

b_n = data sequence

$g(t)$ = pulse waveshape

φ = initial phase (random).

In general, FM spectra are more difficult to calculate than AM spectra because of the highly nonlinear nature of the modulation. A phenomenon is seen in some of the FM spectra, the occurrence of spectral lines, which will occur whenever

$$\omega_d \int_a^b g(t)\ dt = 0,\ \pm\pi,\ \pm 2\pi,\ \ldots$$

where a and b are chosen to include all the energy of the pulse. In general, for any pulse there will be values of the modulation index $k(=\omega_d\ T/\pi)$ at which lines occur in the spectrum. If a pulse shape is used which has no DC component, this integral will be zero, and spectral lines will be present for <u>all</u> values of the modulation index. Most of the signals under consideration have zero DC component, and will, therefore, have lines present in their FM spectra. Sections 2.4.1 and 2.4.2 show FM spectra computed at SIGNATRON by means of

computer simulation, using pseudo-random data, a Hamming window, and a fast Fourier Transform (FFT). The parameters of the simulation are shown in Table 3.

2.4.1 X-Mode FM Spectra

The spectra of the X-mode FM signals are shown in Figs. 14 through 21, as indicated below:

Miller Code (Delay Modulation)

Fig. 14 (16.0 kbps, 9.6 kbps)

Fig. 15 (4.8 kbps, 2.4 kbps)

Paired Selected Ternary (PST)

Fig. 16 (16.0 kbps, 9.6 kbps)

Fig. 17 (4.8 kbps, 2.4 kbps)

Dicode, Twinned Binary

Bipolar (100% Duty Cycle)

These signals have the same spectra. Dicode was the signal simulated.

Fig. 17 (16.0 kbps, 9.6 kbps)

Fig. 18 (4.8 kbps, 2.4 kbps)

Manchester Coding

Conditioned Diphase

These signals have the same spectra. Manchester code was simulated.

Fig. 20 (16.0 kbps, 9.6 kbps)

Fig. 21 (4.8 kbps, 2.4 kbps).

The horizontal axis is the distance from the FM center frequency, in Hz, and the vertical scale is power spectral density in dB (W/Hz). The FM deviation frequency f_d is 8 kHz,

TABLE 3

SIGNAL AND FFT PARAMETERS

Signal Parameters	X-Mode Mod.Index	X-Mode Bit Rate	Quasi Analog
Bit Rates and modulation index	6.67	2400 BPS	75 BPS
	3.33	4800	150
	1.67	9600	300
	1.00	16000	600
			1200
			2400
			4800
Baseband Center Freq.	-		2000 Hz
FM Center Frequency	128,000 Hz		12,000 Hz
Frequency Deviation	±8,000 Hz		±8000 Hz

FFT Parameters

FFT Parameters	X-Mode	Quasi Analog
No. of Points in Transform	1024	1024
Sample Rate (time domain)	819,200/sec	61,440/sec
Record Length	1.25 ms	16.67 ms.
Spacing of Freq.Samples, Δf	800 Hz	60 Hz
Max.Frequency (= Δf x No.of Points)	819,200 Hz	61,440 Hz
Folding Frequency (= Max. Freq/2)	409,600 Hz	30,720

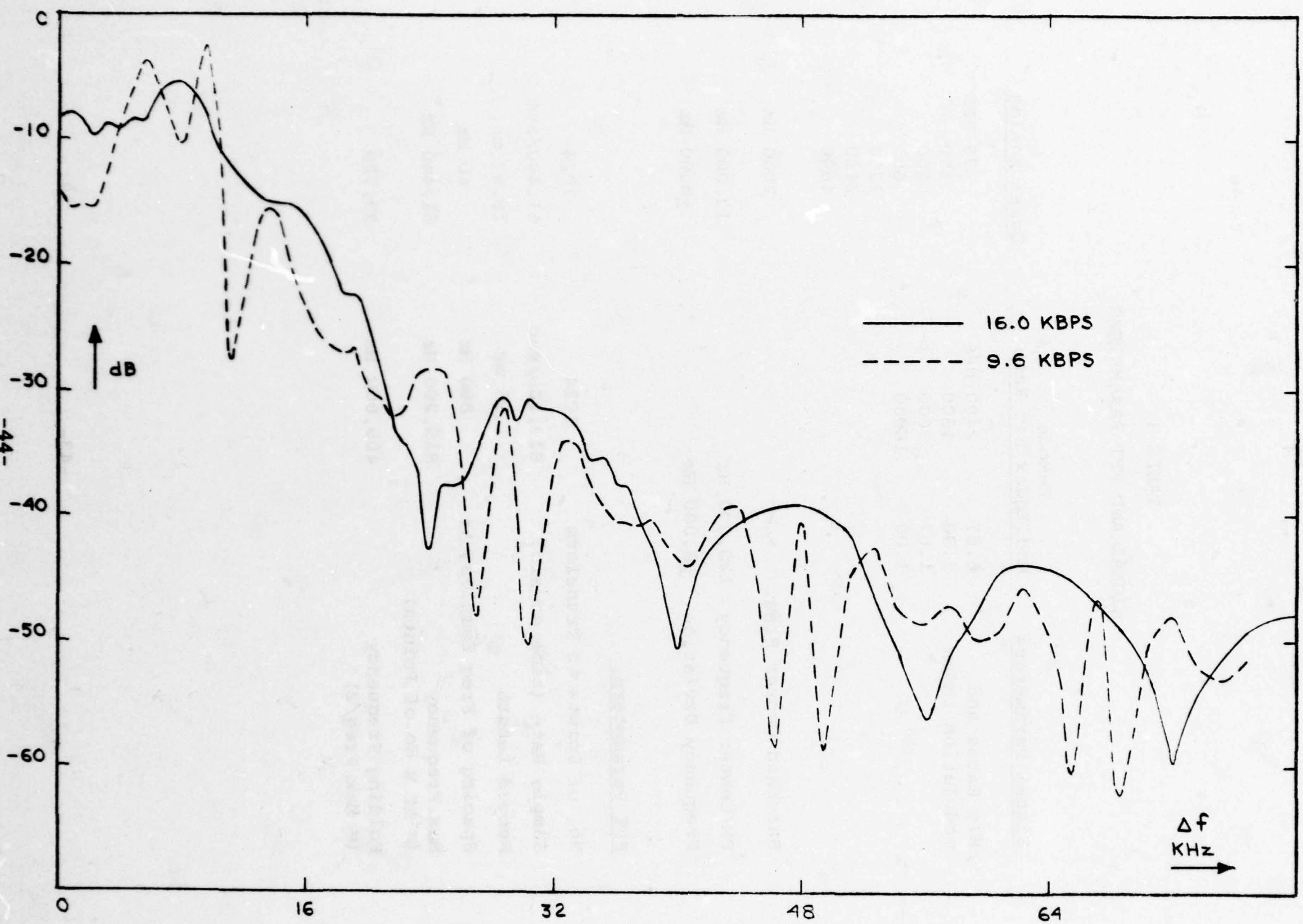


Fig. 14 Miller Code FM Spectrum, 16.0 and 9.6 KBPS

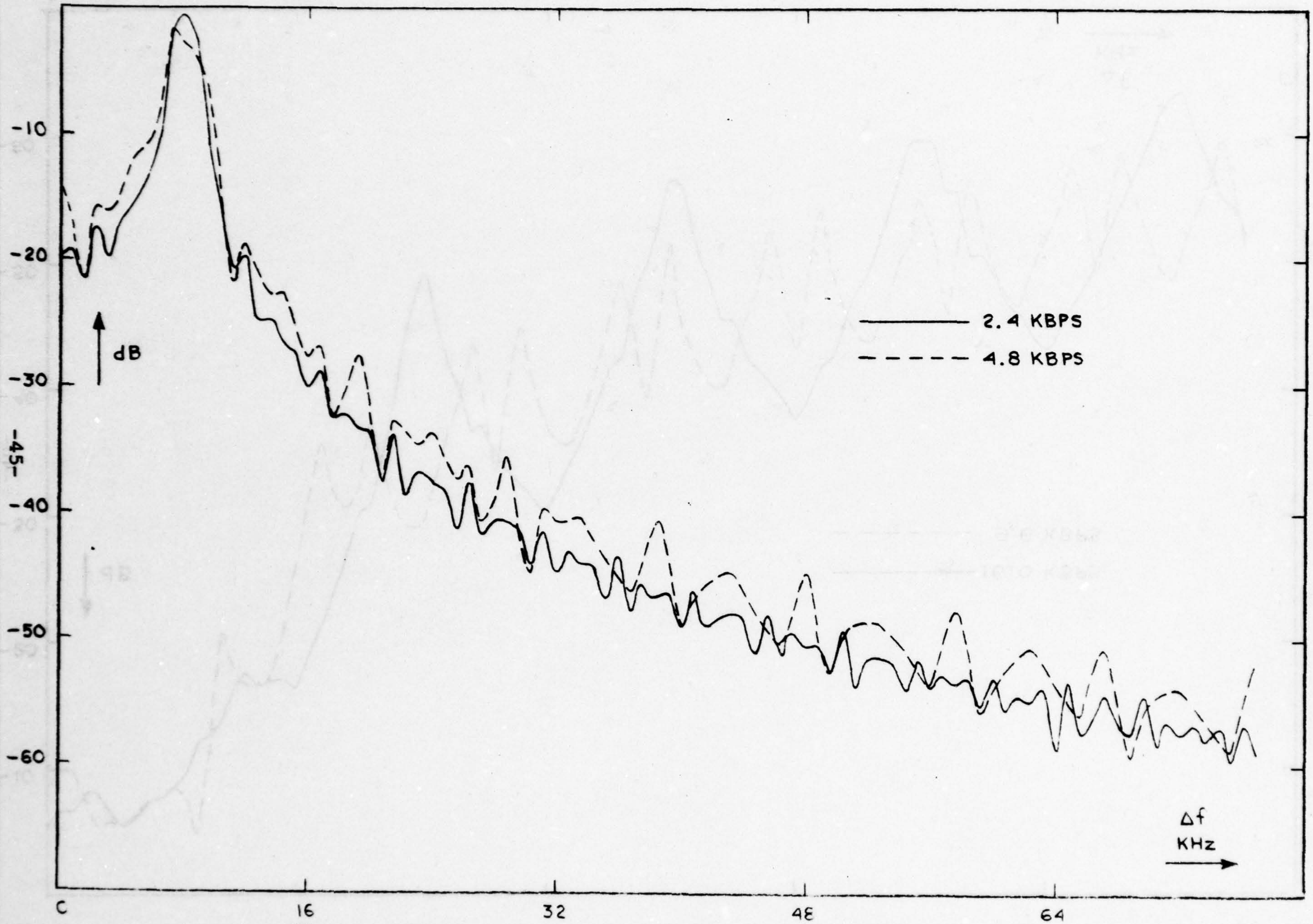


Fig. 15 Miller Code FM Spectrum, 4.8 and 2.4 KBPS

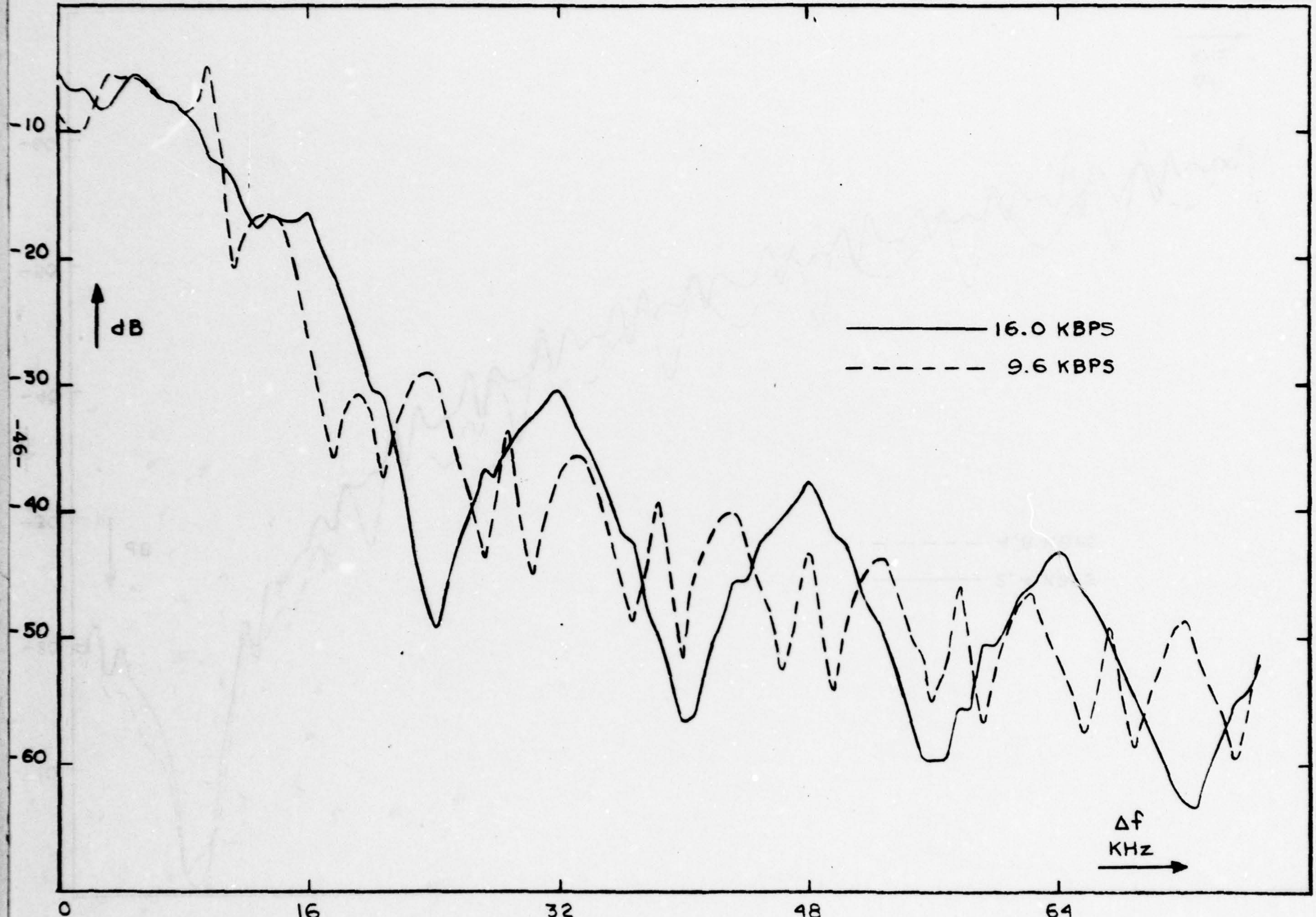


Fig. 16 PST-FM Spectrum, 16.0 and 9.6 KBPS

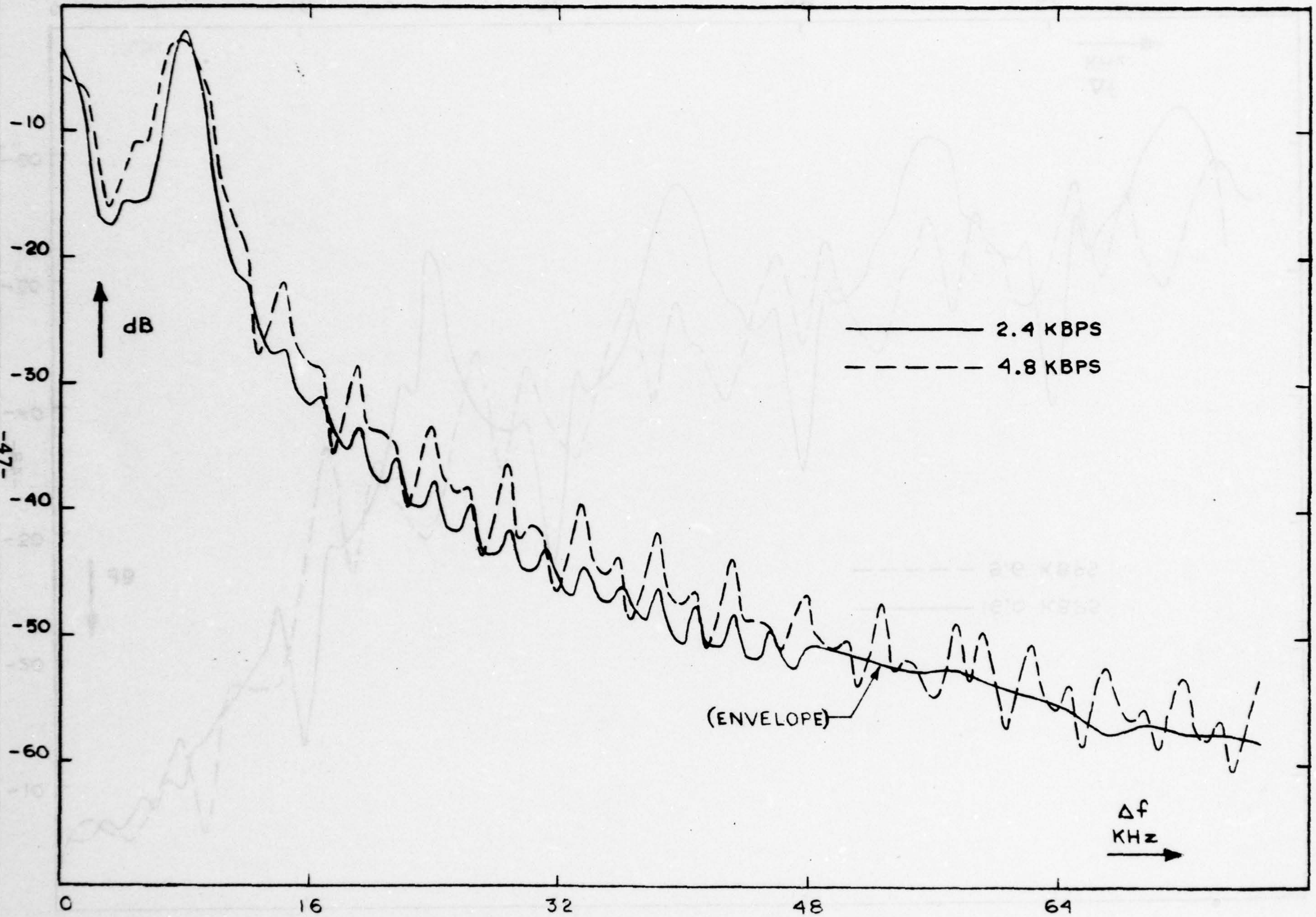


Fig. 17 PST-FM Spectrum, 4.8 and 2.4 KBPS

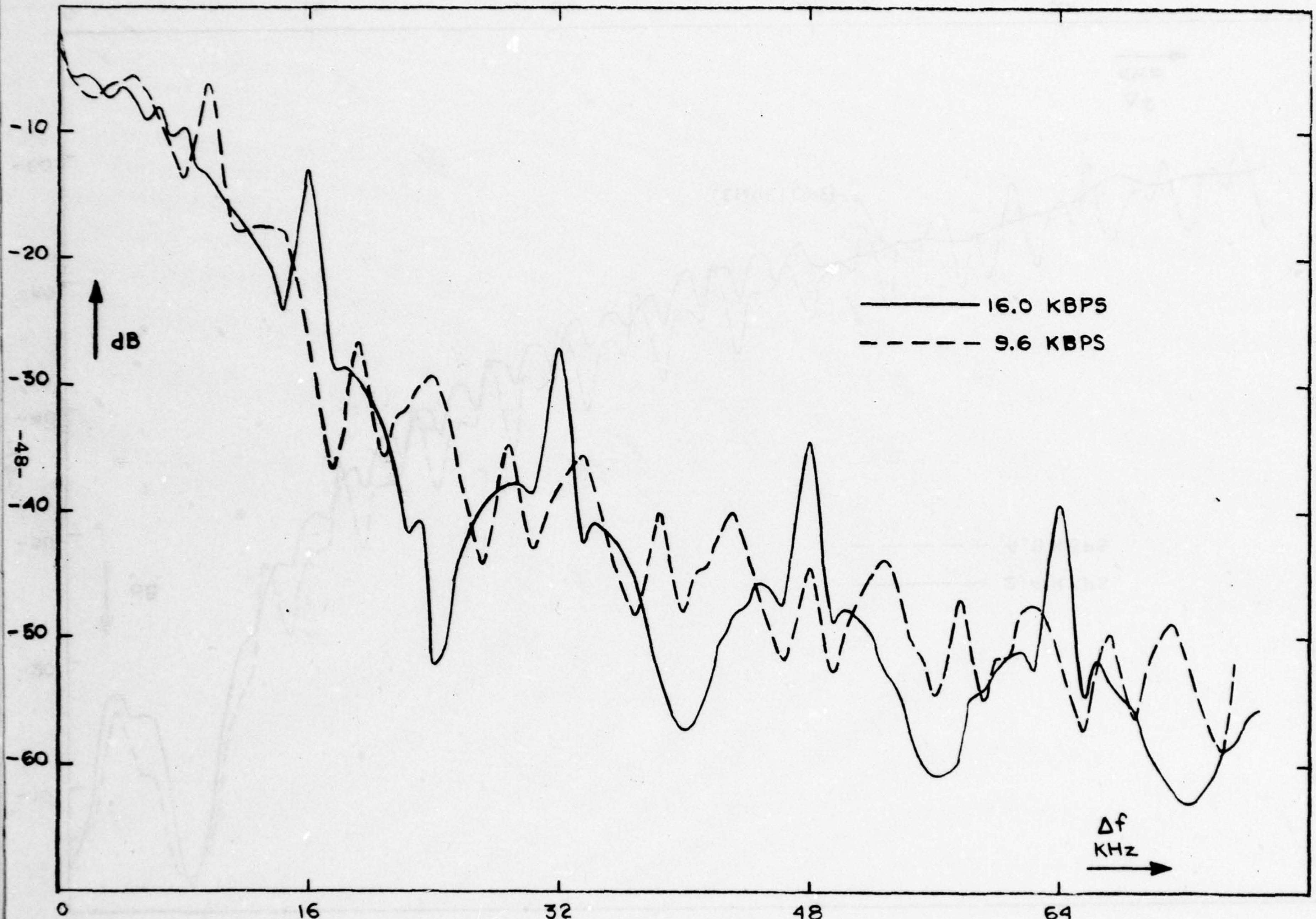


Fig. 18 Dicode FM Spectrum, 16.0 and 9.6 KBPS

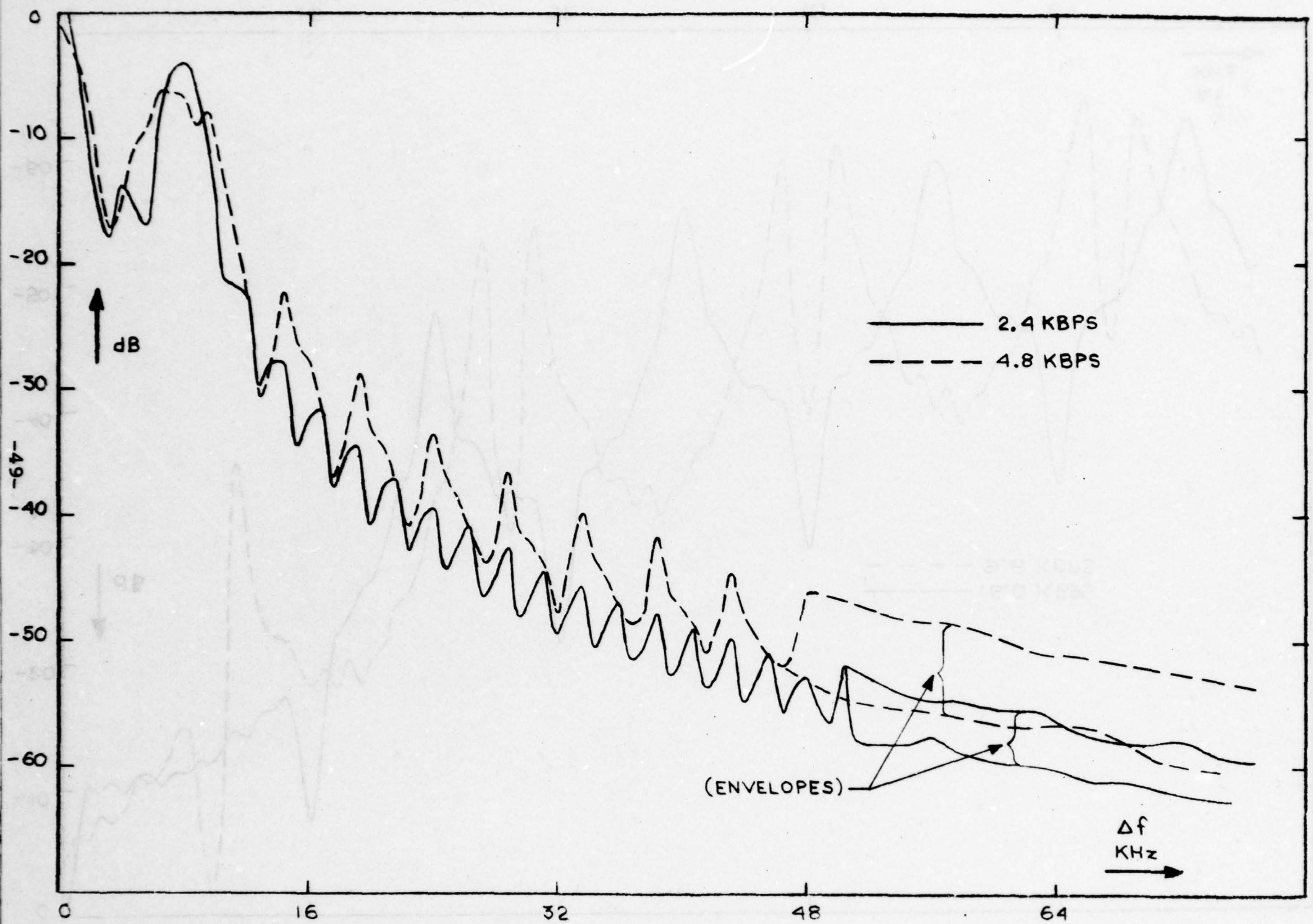


Fig. 19 Dicode FM Spectrum, 4.8 and 2.4 KBPS

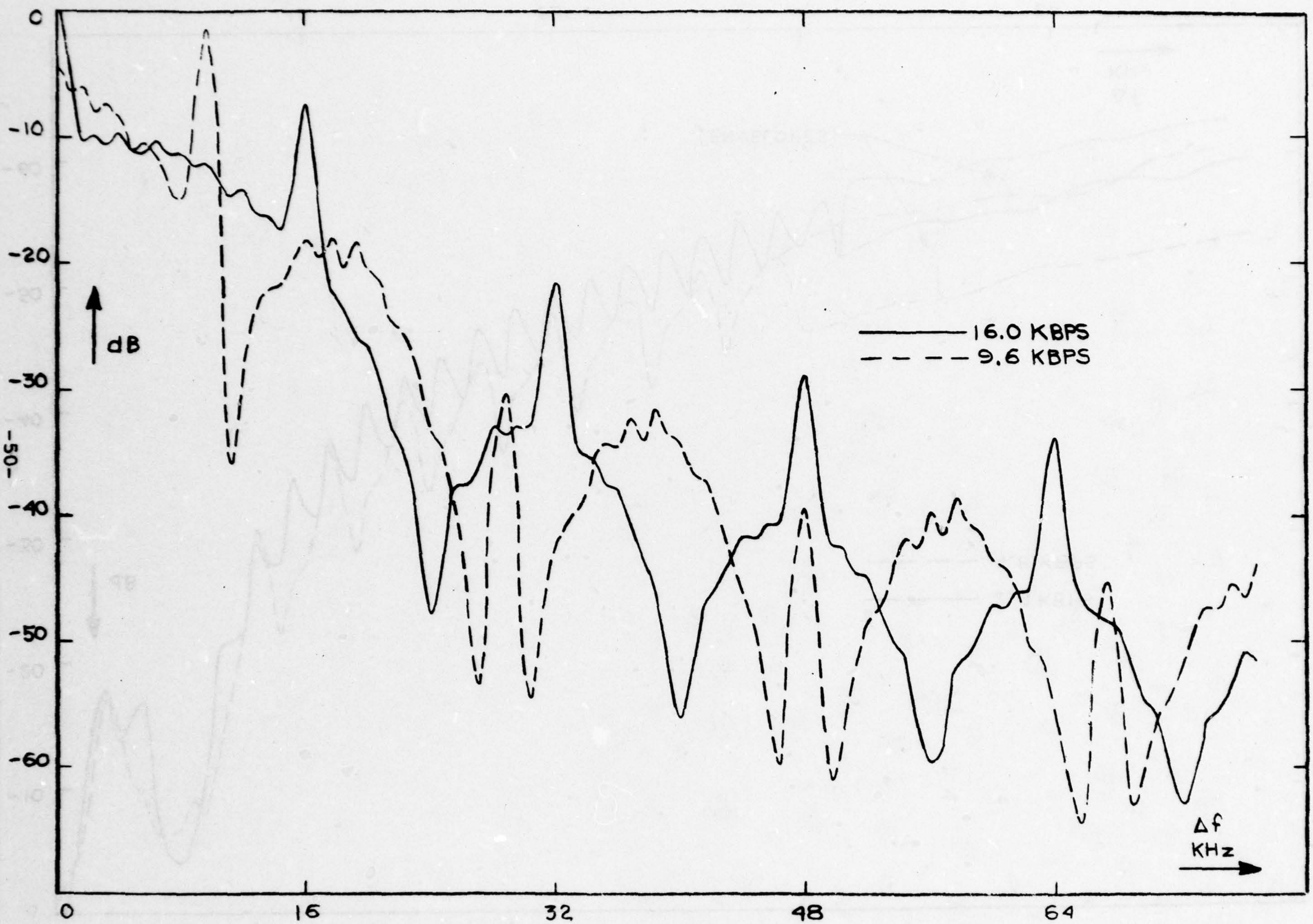


Fig. 20 Manchester Code, Conditioned Diphase FM Spectrum, 16.0 and 9.6 KBPS

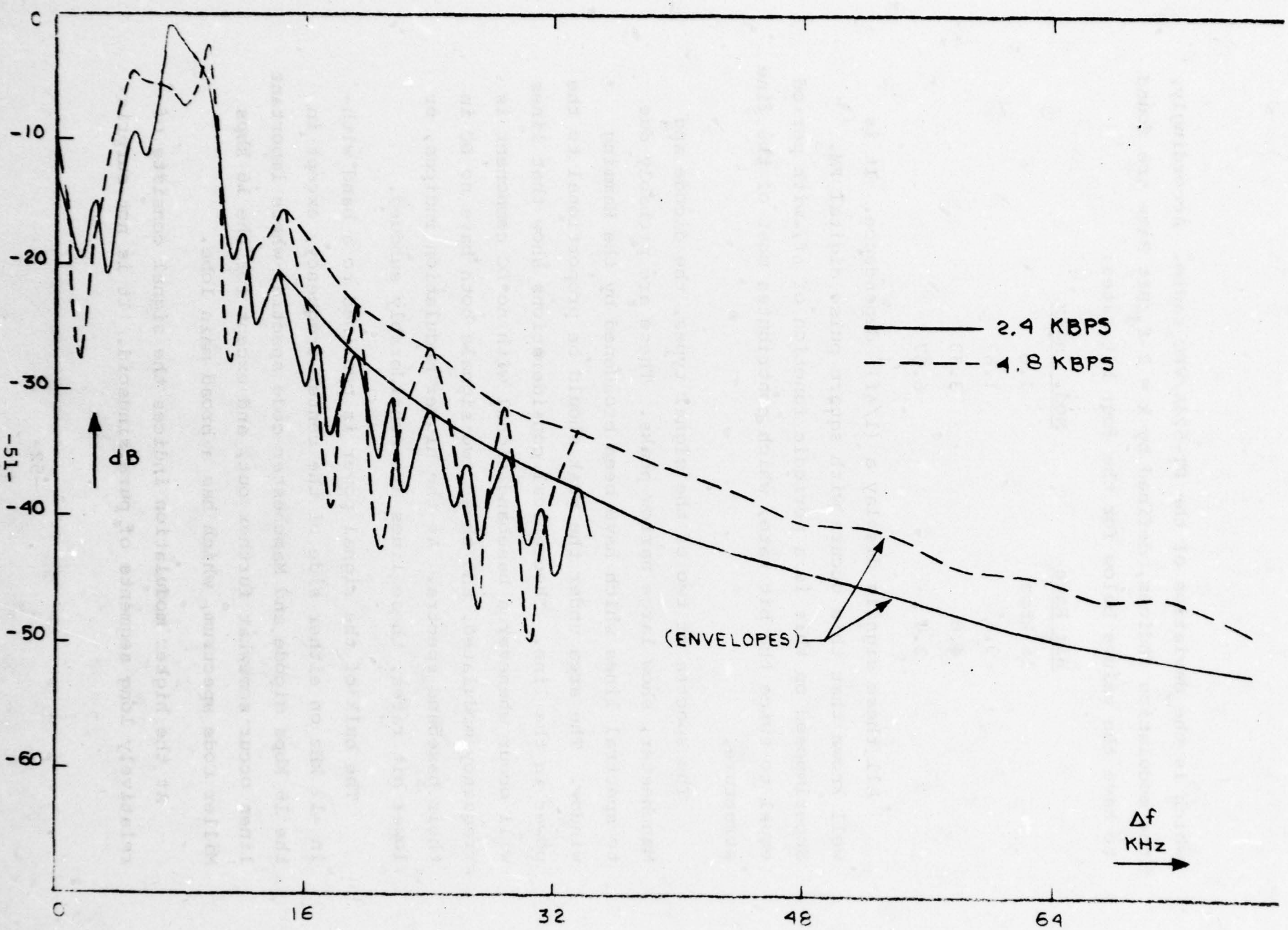


Fig. 21 Manchester Code, Conditioned Diphase FM Spectrum, 4.8 and 2.4 KBPS

which is the deviation of the RT-524A/VRC radio. Accordingly, the modulation indices, defined by $k = 2\ f_d/\text{Bit Rate}$ are found to have the values below for the four bit rates:

Bit Rate	Mod. Index
16 kbps	1.0
9.6	1.67
4.8	3.33
2.4	6.67

All these signals display a $(1/\Delta f)^4$ dependence. It is well known that this occurs with square pulse digital FM. Superimposed on this is a periodic function of Δf, with period equal to twice the bit rate, which contributes most of the fine structure.

The spectra of two of the signal types, the dicode and Manchester, show large narrow peaks. These are probably due to spectral lines which have been broadened by the Hamming window. The area under the peak should be proportional to the power in the line. Theoretical considerations show that lines will occur whenever a baseband signal with no DC component is frequency modulated, and these two signals both have no DC in their baseband spectra. At the higher modulation indices, or lower bit rates, these lines are considerably subdued.

The bulk of the signal power is confined to a band within ±12 kHz on either side of the center frequency, except in the 16 kbps dicode and Manchester code spectra, where important lines occur somewhat further out, and except for the 16 kbps Miller Code spectrum, which has a broad main lobe.

At the higher modulation indices the signal consists of relatively long segments of pure sinusoid. It is not surpris-

ing that the spectrum has peaks at or near the frequencies of these tones; f_c ±8 kHz for the binary signals, Miller and Manchester; and f_c, f_c ±8 kHz for the ternary signals, Dicode and PST.

2.4.2 Quasi-Analog FM Spectra

The spectra of the quasi analog FM signals are given in Figs. 22 and 23. Figure 22 shows MSK-FM and Fig. 23 shows FSK-FM with an FSK modulation index of 1. Since the generation of the signal requires two stages of frequency modulation we will describe it in some detail to avoid any confusion about the parameters. The initial signal is a train of positive or negative NRZ pulses. This is fed into the first modulator to produce a continuous phase FSK signal. The modulation index is defined to be $k = 2\ f_d/(\text{Bit rate})$, where the two tone frequencies are $f_c \pm f_d$. The RT-524A/VRC radio has an audio passband from 500 Hz to 3500 Hz, so the center frequency chosen for the first modulation is $f_c = 2000$ Hz. If $k = 0.5$, this signal is minimum shift keying, MSK. At other modulation indices it is called FSK or CPFSK.

This signal is applied to the second modulator which produces the final signal. The deviation frequency of the radio is 8 kHz, so this same value was used in the simulation. Notice that the input to the second modulator is a segment of a sinusoid, so that during each single bit, its output has sinusoidally varying frequency.

The spectra in Fig. 22 are for MSK-FM; that is, the modulation index in the first modulation is 0.5. The curves shown are for 75, 600, and 4800 bps. All three have envelopes of roughly the same shape. The fine structure changes consider-

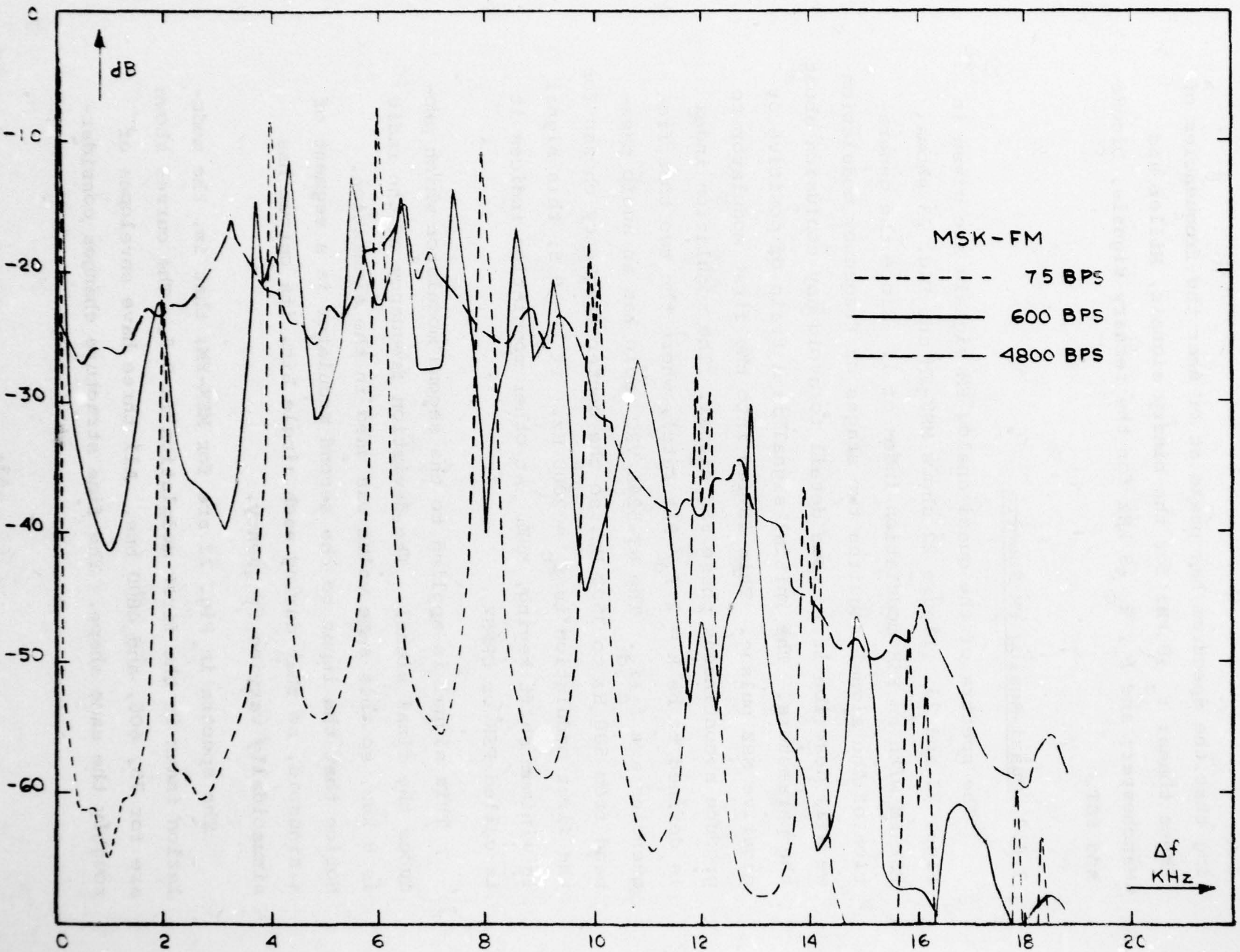


Fig. 22 MSK-FM Spectrum

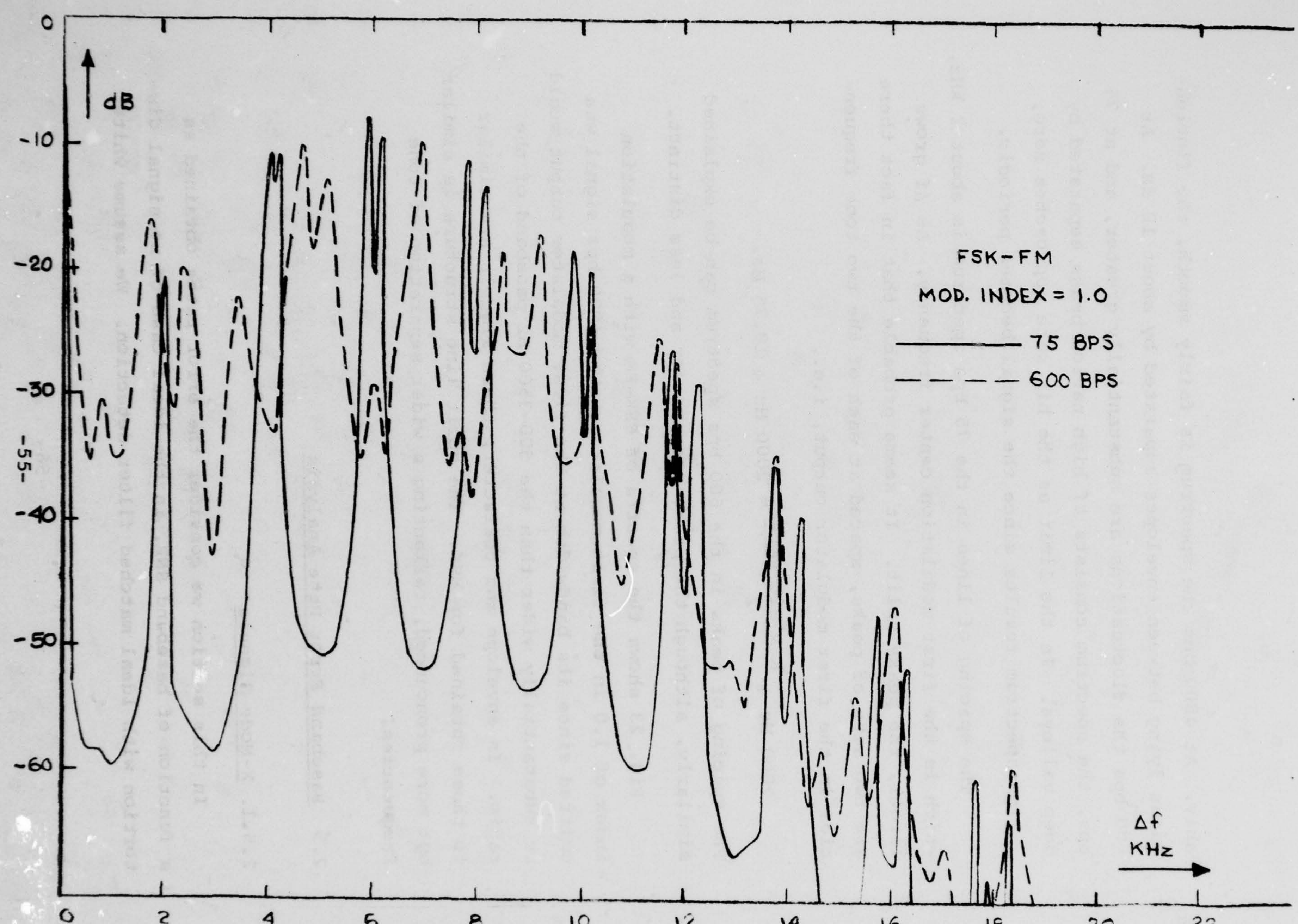


Fig. 23 FSK-FM Spectrum, Mod. Index = 1.0

ably. At 4800 bps the spectrum is fairly smooth, the fluctuations lying between envelopes separated by about 10 dB. At 600 bps the fluctuations are substantially greater, and at 75 bps, the spectrum consists of high narrow peaks separated by deep valleys. In the limit as the bit rate approaches zero, a line spectrum results since the signal becomes periodic.

The spacing of lines in the 75 bps spectrum is about 2 kHz, which is the first modulation center frequency. As Δf grows larger, the peaks split. It seems probable that in fact there are two sets of peaks, spaced at each of the two tone frequencies in the first modulator output, i.e.,

$$2000 \text{ Hz} \pm \frac{\text{k x Bit Rate}}{2} = 2000 \text{ Hz} \pm 18.75 \text{ Hz}.$$

The spacing of peaks in the 600 bps spectrum can be explained similarly, although the peaks are broader and less distinct.

Fig. 23 shows the spectra of FSK-FM with a modulation index of 1.0 in the first modulator. The 4800 bps signal was omitted since its bandwidth at the first modulator output would be substantially wider than the 500-3500 Hz passband of the radio. In envelope and character, these signals are similar to those obtained for MSK. The split line structure is similar but more pronounced, reflecting a wider separation of tone frequencies.

2.5 Baseband Error Rate Analysis

2.5.1 X-Mode Signals

In this section we consider the error rates obtained as a function of baseband SNR, in the ideal case of no signal distortion with ideal matched filter detection. We assume white

Gaussian noise. For this case, the filter output has a signal-to-noise ratio equal to $2E_b/N_0$, where E_b is the peak energy of a bit (i.e., the energy of a 1 bit or a -1 bit, but not the energy of a zero bit in the ternary systems) and where N_0 is the single sided noise spectral density. This is twice the usual formula for bandpass systems, because bandpass systems admit both upper and lower sideband noise.

As a starting point we calculate the error rate for an NRZ signal. For this case, an error will occur if the noise plus signal has polarity opposite that of the signal; in other words, if the noise has larger magnitude than the signal, and opposite polarity. Taking the signal voltage as $\sqrt{E_b}$, the rms noise voltage will be $\sqrt{N_0/2}$, and the error probability is given by

$$P_e = \int_{\sqrt{E_b}}^{\infty} \frac{1}{\sqrt{2\pi}\,\sqrt{N_o/2}} \exp\left\{- x^2/2(N_o/2)\right\} dx$$

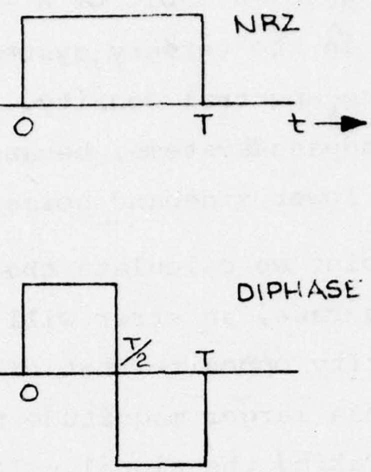
$$= \frac{1}{2} \operatorname{erfc} \sqrt{E_b/N_o} = \frac{1}{2} \operatorname{erfc} \sqrt{\gamma}$$

where

$$\gamma = E_b/N_o.$$

The same error rate will apply to the diphase pulse, when it is match-filtered, because both pulses, NRZ and diphase, have the same energy. The two pulse shapes, shown in Fig. 11, are repeated in the illustration on the following page.

This will be the probability of error in detecting a bit for the two signals using antipodal pulses, i.e., Manchester Code and Conditioned Diphase. The ternary signals will only



NRZ and Diphase Pulses

have half as much noise margin. Therefore there will be a 6 dB loss, and the basic bit detection error rate will be $\frac{1}{2}$ erfc $(\sqrt{\gamma}/2)$. This rate, denoted as P_A, applies when a 0 is detected as a 1 or as a -1. When a 1 or -1 is transmitted, the errors can be divided into two categories, the more common errors when a 1 or -1 is detected as zero, and the rarer errors when a 1 or -1 is detected as a -1 or 1 respectively. The second type of error will be quite rare, occurring at the rate $P_C = \frac{1}{2}$ erfc $(3\sqrt{\gamma}/2)$. Finally the more common error of detecting a ± 1 as a 0 has the rate $P_B = P_A - P_C$.

Because the various signals have more or less complicated rules for encoding and decoding, the actual error rate in the decoded data is not necessarily just the detection error rate, due to error propagation effects. Accordingly for each signal we will give the decoding rules and determine the decoded

error rate. In some of the signals, because of dependence between adjacent symbols or groups of symbols, it is theoretically possible to reduce the error rate by sophisticated techniques, such as the use of soft detection decisions followed by some sort of maximum likelihood search procedure any time an error is detected. We will assume that such procedures are not used, but rather that the decoding is done as simply as possible. We will also assume that detection errors are sufficiently rare that the analysis may be confined to single errors only.

Our notation is as follows

a_n input data stream (binary, ± 1)
b_n coded data stream
c_n detected data stream
d_n decoded data stream (binary, ± 1)

Dicode: This is encoded by the rule

$$b_n = \tfrac{1}{2}(a_n + a_{n-1}).$$

The decoding rule is given by the table:

if	$c_n = 1$	then $d_n = 1$
	$c_n = -1$	$d_n = -1$
	$c_n = 0$	$d_n = c_{n-1}$

There are six possible detection errors

a)	$b_n = 1$	$c_n = 0$	$P(c_n \mid b_n) = P_B$
b)	1	-1	P_C
c)	0	1	P_A
d)	0	-1	P_A

e)	-1	1	P_C
f)	-1	0	P_B

Because of symmetry, the following pairs will have the same effect on the error rate: (a,f), (b,e), (c,d). Types a,b,e, and f will initiate a sequence of errors, starting with d_n and continuing until $c_{n+i} = \pm 1$ occurs. For random data, one half the c_n's will be zeros. Accordingly the stream of errors will have length 1 with probability ½, length 2 with probability ¼, and length n with probability $1/2^n$. The expected length of the error stream is given by

$$E[n] = \sum_{n=1}^{\infty} n/2^n = 2.$$

Errors of type c) and d) will lead to no decoded error one half the time, and will initiate a stream of errors one half the time. The expected length in that case is also 2. Weighting each error stream by the probability of b_n and the probability of the detection error, and summing we obtain the total probability of a decoded bit error:

	$P(b_n)$	P(detection error)	E(N)	Product
a	1/4	P_B	2	$P_B/2$
b	1/4	P_C	2	$P_C/2$
c	1/2	P_A	½ x 2	$P_A/2$
d	1/2	P_A	½ x 2	$P_A/2$
e	1/4	P_C	2	$P_C/2$
f	1/4	P_B	2	$P_B/2$
				$P_A + P_B + P_C$

Since $P_A = P_B + P_C$, we have

$$P_e = 2\,P_A = \text{erfc}\left(\frac{\sqrt{\gamma}}{2}\right).$$

<u>Bipolar</u>. This is encoded according to the rule:

$a_n = -1 \qquad b_n = 0$

$a_n = 1 \qquad b_n = \pm 1$, strictly alternating.

The decoding rule is simply:

$c_n = 0 \qquad d_n = -1$

$c_n = \pm 1 \qquad d_n = 1.$

Since no memory is involved in the decoding rule, no error propagation will occur. The same six detection errors can occur as in dicode. Types 2, c, d, and f will each produce a single decoded bit in error, while b and e will not produce a decoding error. The total error rate is computed in the following chart.

	$P(b_n)$	P(detection error)	
a	1/4	P_B	$P_B/4$
c	1/2	P_A	$P_A/2$
d	1/2	P_A	$P_A/2$
f	1/4	P_B	$P_B/4$
			$P_A + P_B/2$

Since $P_A \approx P_B$,

$$P_e = \frac{3}{2} P_A = \frac{3}{4} \operatorname{erfc}\left(\frac{\sqrt{\gamma}}{2}\right).$$

Paired Selected Ternary. This is encoded according to the table below:

a_{2n-1}	a_{2n}	b_{2n-1}	b_{2n}
-1	-1	-1	1
-1	1	0	±1
1	-1	±1	0
1	1	1	-1

The occurrence of either of the symbols with ± signs is strictly alternating.

The decoding rule is simply the same chart rearranged:

c_{2n-1}	c_{2n}	d_{2n-1}	d_{2n}
-1	1	-1	-1
0	±1	-1	1
±1	0	1	-1
1	-1	1	1

with the pairs (0,0) (-1,-1) and (1,1) illegal. The occurrence of one of these indicates either an out-of-sync condition or else an error. When the out-of-sync condition occurs, with a random input stream, fully 6/16 of the pairs (31%) will have one of the illegal character pairs, so that resynchronization should be done only when frequent occurrence takes place

over some span of time. Streams of all 1's, all -1's and strictly alternating 1's and -1's, when coded, contain no illegal pairs. For this reason they are undesirable as "idle" signals between messages, since idle signals should permit detection of synchronization errors, if any.

The error rate for this signal is calculated by the same method as in bipolar, but taking pairs of symbols instead. Due to the large number of cases, we will omit most of the details, except for one point. If one of the illegal pairs occurs we may either accept both bits as errors, or else fill in with arbitrary bits, which have 50% probability of being right. In the first case the error rate is

$$P_e = \frac{7}{8} \operatorname{erfc} \left(\frac{\sqrt{\gamma}}{2}\right)$$

and in the second case it is reduced to

$$P_e = \frac{5}{8} \operatorname{erfc} \left(\frac{\sqrt{\gamma}}{2}\right) .$$

Manchester Coding. This is similar to NRZ except the diphase pulse which has a sign change at mid-bit is used instead of the NRZ pulse. No coding is used. The error rate is the same as for NRZ pulses, $Pe = \frac{1}{2} \operatorname{erfc} (\sqrt{\gamma})$

Conditioned Diphase. This uses the diphase pulse, but the input data is precoded,

$$b_n = b_{n-1} \oplus a_n$$

where $\oplus$ means exclusive-or. Decoding is done by the formula

$$d_n = c_n \oplus c_{n-1}.$$

Since each of the detected digits (the c's) affects two of the decoded digits, the decoded error rate will be twice the detec-

tion error rate,

$$P_e = \text{erfc}\,(\sqrt{\gamma}).$$

Miller Coding (Delay Modulation) uses NRZ pulses to represent -1's and diphase pulses to represent +1's, with the polarities chosen so that each interval between sign changes has length of T, 1½T, or 2T. This is accomplished by the following rule: If adjacent pulses are -11, 11, or 1-1, the new symbol start with the same polarity as the last half of the previous symbol. With the pair -1-1, the new symbol has polarity opposite that of the previous symbol.

This signal may be detected by having two equal gain matched filters, one matched to an NRZ pulse, and the other matched to a diphase pulse. At each sampling instant the filter outputs are compared and the one having the larger magnitude is selected. If it is the NRZ filter, the symbol is decoded as -1; if the diphase filter output has the larger magnitude the symbol is decoded as +1. Let X_N and X_D be the NRZ and diphase filter outputs, respectively. Suppose that there is a positive NRZ pulse. Then $X_n = k\left(\sqrt{E_b} + n_N\right)$, and $X_D = kn_D$. Here k is the gain of filters. Because the filters are orthogonal and have equal gain, the noise samples X_N and X_D will be independent and identically distributed, and because of our earlier white Gaussian hypothesis, n_N and n_D will be zero mean Gaussian, $E(n_D^{\,2}) = E(n_N^{\,2}) = N_0/2$. A detection error will occur if $|X_D| > |X_N|$. Figure 24 shows the error region, and contours of the joint probability density of X_N and X_D. The probability of this error region can be expressed in terms of a one dimensional integral, since the lines bounding the region are perpendicular. It is found that

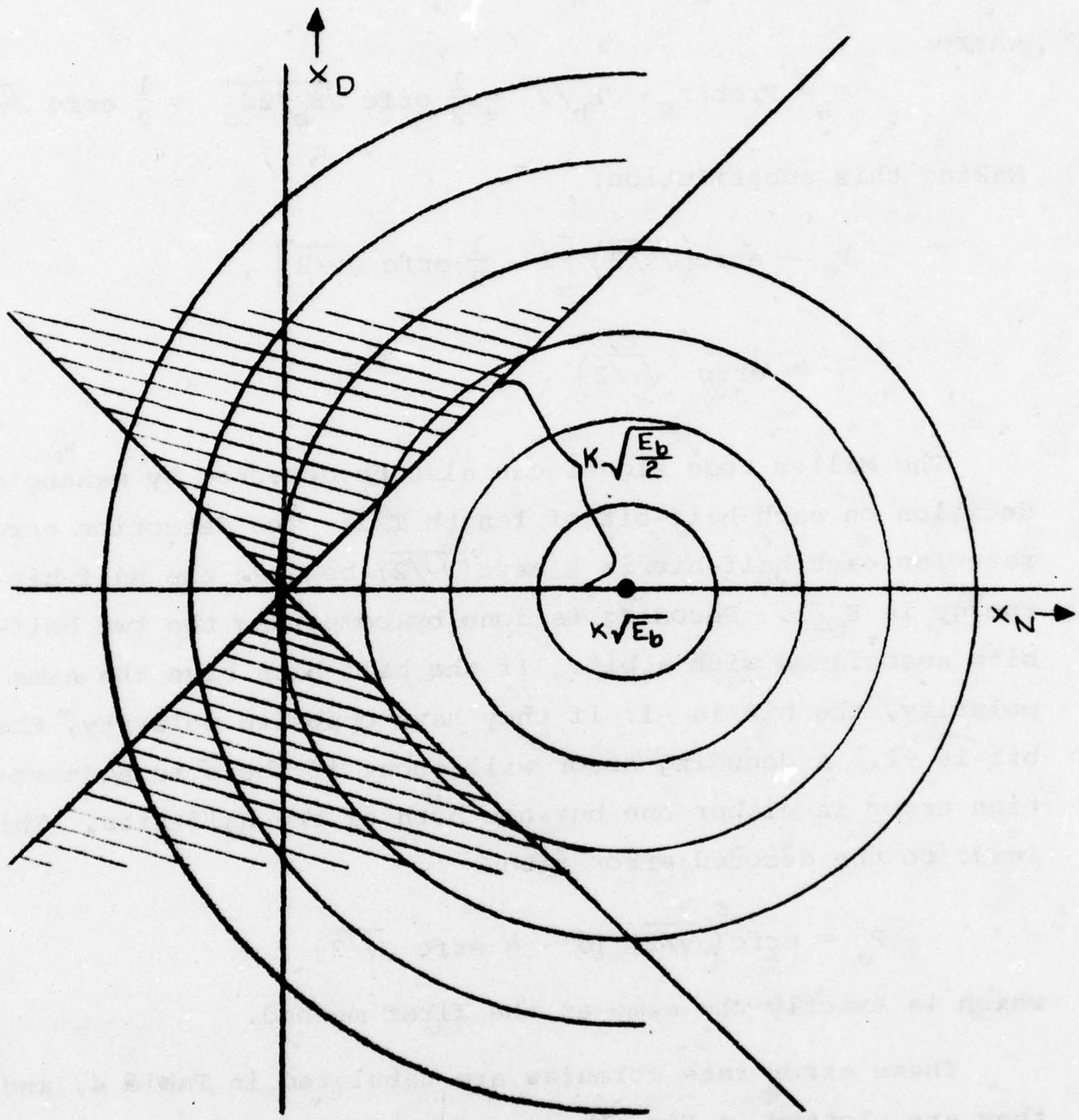


Fig. 24 Error Region (Shaded) for Miller Code Detection

$$P_e = 2\,P_A(1 - P_A)$$

where

$$P_A = \mathrm{Prob}\left(n_N > \sqrt{E_b/2}\right) = \frac{1}{2}\,\mathrm{erfc}\,\sqrt{E_b/2N_o} = \frac{1}{2}\,\mathrm{erfc}\,\sqrt{\gamma/2}\;.$$

Making this substitution,

$$P_e = \mathrm{erfc}\left(\sqrt{\gamma/2}\right)\left[1 - \frac{1}{2}\,\mathrm{erfc}\,\sqrt{\gamma/2}\right]$$

$$\approx \mathrm{erfc}\left(\sqrt{\gamma/2}\right)\;.$$

The Miller Code signal can also be detected by making a decision on each half-bit of length $T/2$. The detection error rate for each half-bit is $\frac{1}{2}\,\mathrm{erfc}(\sqrt{\gamma/2})$ because the half-bit energy is $E_b/2$. Decoding is done by comparing the two half-bits associated with a bit. If the half-bits have the same polarity, the bit is -1; if they have opposite polarity, the bit is +1. A decoding error will occur if there is a detection error in either one but not both of the half-bits. This leads to the decoded error rate

$$P_e = \mathrm{erfc}(\sqrt{\gamma/2})\;[1 - \tfrac{1}{2}\,\mathrm{erfc}\,\sqrt{\gamma/2}]\;,$$

which is exactly the same as the first method.

These error rate formulas are tabulated in Table 4, and they are plotted in Fig. 25 as a function of SNR.

2.5.2 Quasi-Analog Signals

The error rates for quasi-analog signals have been given in Section 2.2, namely

Table 4 X-Mode Error Rates

Dicode	$\mathrm{erfc}\left(\frac{\sqrt{\gamma}}{2}\right)$
Bipolar	$\frac{3}{4}\,\mathrm{erfc}\left(\frac{\sqrt{\gamma}}{2}\right)$
Paired Selected Ternary	$\frac{5}{8}\,\mathrm{erfc}\left(\frac{\sqrt{\gamma}}{2}\right)$ detected errors filled in
	$\frac{7}{8}\,\mathrm{erfc}\left(\frac{\sqrt{\gamma}}{2}\right)$ detected errors not filled in.
Manchester Coding	$\frac{1}{2}\,\mathrm{erfc}(\sqrt{\gamma})$
Conditioned Diphase	$\mathrm{erfc}\ (\sqrt{\gamma})$
Miller Coding	$\mathrm{erfc}\left(\sqrt{\frac{\gamma}{2}}\right)$

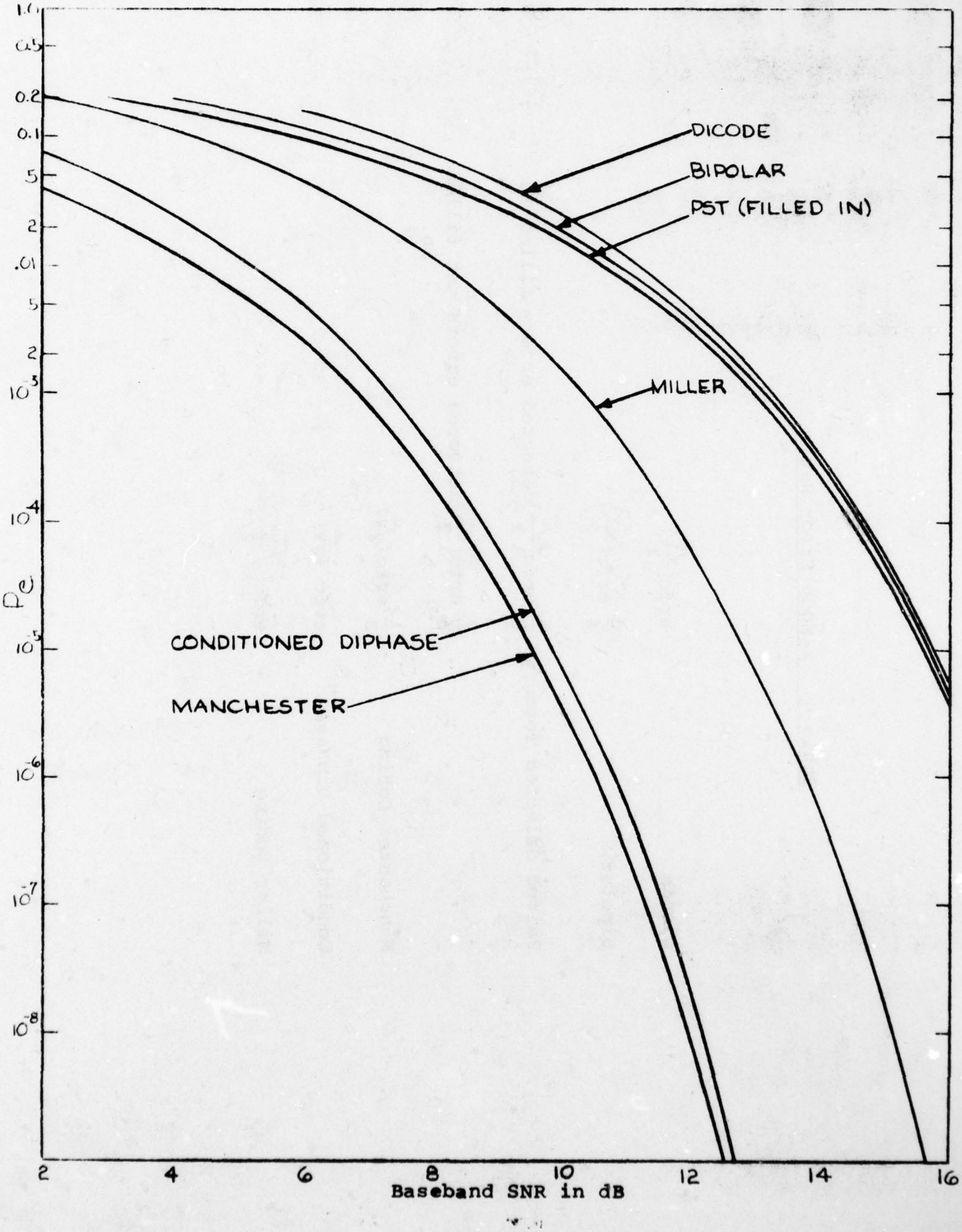


Fig. 25 X-Mode Waveform Error Rates

$$P_e = \begin{cases} \frac{1}{4} e^{-\rho} & \text{Binary FSK} \\ \dfrac{3}{4\sqrt{2\pi}} \dfrac{\operatorname{ctn}(\pi/8)}{\sqrt{\rho}\cos\pi/4} e^{-2\rho \sin^2(\pi/8)} & \text{4-LFSK} \\ \frac{1}{2} \operatorname{erfc} \sqrt{\rho} & \text{MSK.} \end{cases}$$

Simulations have shown that the theoretical error rate for FSK can not be achieved, due to filter effects. Figure 26 shows four error rate curves. Two of them are the theoretical curves for binary FSK and MSK. The curve marked "Binary FSK (Simulation)" is a composite curve, taken from the technical literature. The 4 LFSK error rate curve is the result of degrading the Binary FSK Simulation curve by 4.5 dB, which is the asymptotic difference between binary and 4 LFSK performance.

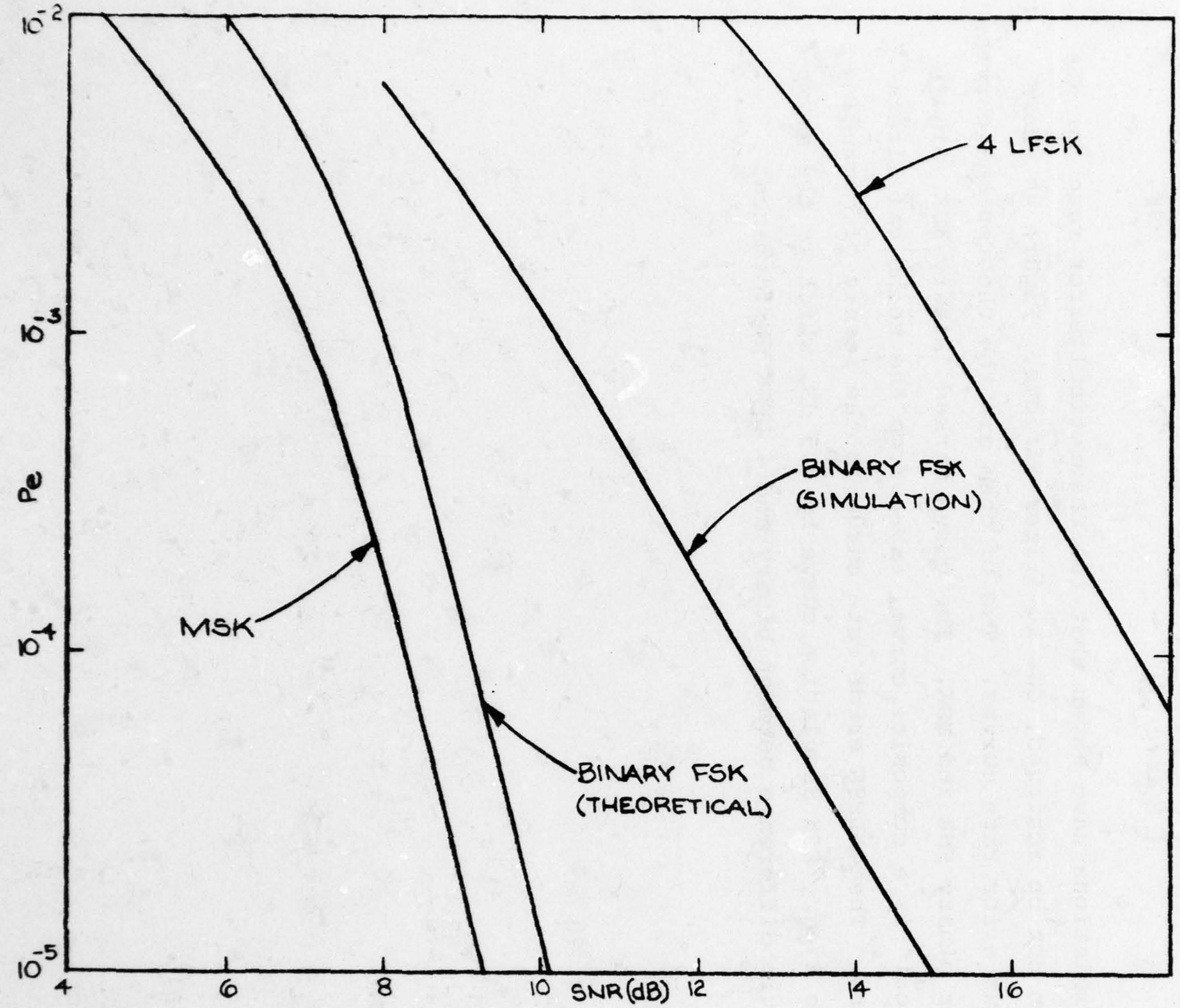


Fig. 26 Quasi-Analog Mode Error Rates

References for Section 2

Lucky, R.W., Salz, J., Weldon, E.J. Jr., 1968, Principles of Data Communication, McGraw-Hill Book Co., New York, N.Y. 1968.

Mazo, J.E., Salz, J., 1966, "Theory of Error Rate for Digital FM", BSTJ, Vol. 45, Nov. 1966, pp. 1511-1535.

SECTION III

IMPLEMENTATION

In this section we describe the implementation of the Baseband Signal Converter. This section is an abridged version of the Equipment Description Manual, which contains full schematic drawings of the Converter.

3.1 Card Breakdown

The Signal Converters each consist of eight wire wrap cards. The cards have been given the following designations:

1. XSG - X mode Signal Generator,
2. QAR - Quasi Analog Signal and Rate Generator,
3. MSR - MSK Carrier and Clock Recovery Loops,
4. ICO - Integrator Combiner,
5. FDS - FM Discriminator and SSB Converter,
6. TTI - Time Tracker Filter and I/O,
7. SSI - Squelch and Sync Inhibit,
8. TCL - Timing and Control Logic.

3.1.1 Philosophy of Receiver Structures

We have attempted in all cases to design receiver structures that are optimum for the non-bandlimited white noise case. For the case of FSK modulation this is not strictly possible because the optimum receiver structure is a function of the modulation index. It is known that a discriminator preceded by a filter approximately equal to the data rate can achieve performance within 1 dB of the optimum achievable with any frequency discriminator. Therefore a highly accurate baseband discriminator has

been designed. The discriminator is preceded by a variable bandwidth filter which can be set from the front panel. The modulation index is variable over a range of 0.2 to 1.9 in steps of 0.1.

3.1.2 System Block Diagram

Figure 27 is a block diagram of the signal converter system. The blocks show the functional elements and their interconnection. The location of the circuitry which performs each function is shown by the card designations in the corner of each block.

Section 3.2 discusses the design of the system in detail. Some discussion of the design choices is given below.

X-Mode Signal Generator

This functional unit is also a physical unit in that it is constructed on one card which produces all codes. The card is all digital except for the D to A to converter. It is very accurate and drift free. The clock generation for the various data rates is also performed on this card.

QA-Mode Signal Generator

This card is also both a functional and physical unit. The waveforms are generated by digital techniques. This card is also all digital except for the D to A converter. The oscillator driving this card is also the timing source for the X-mode generators, and is used to produce some LO's needed in the receiver portion of the signal converter.

Single Sideband Converter

As the QA mode received signals cannot be processed at the 2 kHz IF frequency, a translation to 12 kHz is performed.

The LO's in all cards are derived from a highly stable source so frequency errors are very small. This is necessary to detect the low data rate, low modulation index signals.

FSK Receiver

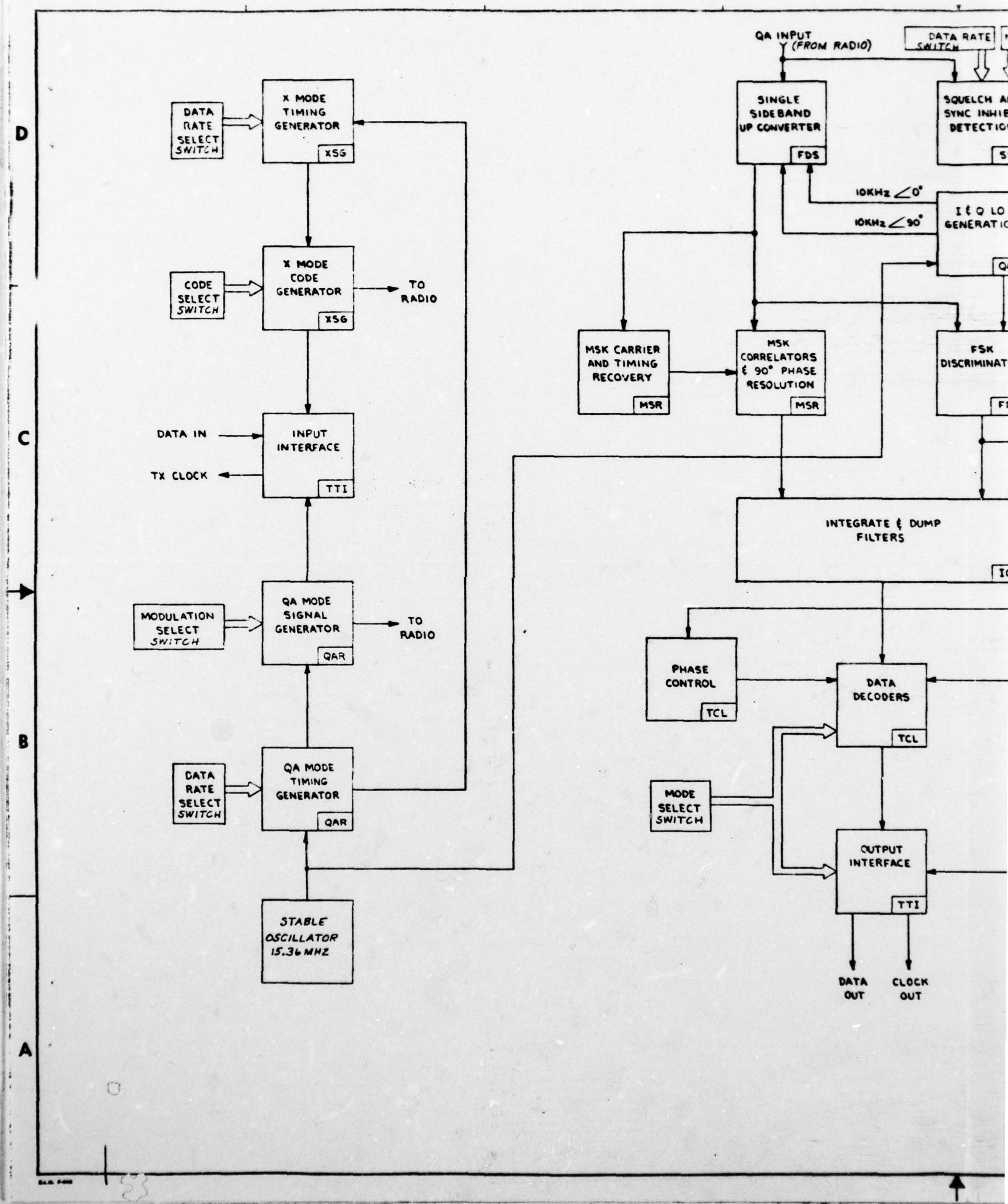
A well designed baseband discriminator having a selectable predetection filter is used as the FSK receiver. This type of receiver was selected because it performs well over a wide range of data rates.

MSK Receiver

The MSK receiver is the most complex of all the receivers. It has two PLL's with extremely narrow bandwidths at the data rates involved. The clock for this receiver is produced from these two loops, as well as the reference waveforms required to perform the matched filtering required in this receiver. The modulator is the same as is used with FSK, except that the modulations index is forced to 0.5 and the data is differentially encoded.

Integrator Network

All received signals in all modes are integrated over some period or periods before the data decision is made. A single pair of integrators is used in all cases. This circuit is designed using highly stable op amps which require no alignment for laboratory use. The integrator network also forms part of the time discriminator used for clock recovery in all modes except MSK.


D
C
B
A
DATA RATE SELECT SWITCH
X MODE TIMING GENERATOR
XSG
CODE SELECT SWITCH
X MODE CODE GENERATOR
XSG
TO RADIO
DATA IN
TX CLOCK
INPUT INTERFACE
TTI
MODULATION SELECT SWITCH
QA MODE SIGNAL GENERATOR
QAR
TO RADIO
DATA RATE SELECT SWITCH
QA MODE TIMING GENERATOR
QAR
STABLE OSCILLATOR 15.36 MHZ
QA INPUT (FROM RADIO)
DATA RATE SWITCH
SINGLE SIDEBAND UP CONVERTER
FDS
SQUELCH A
SYNC INHIB
DETECTIO
10KHz ∠0°
10KHz ∠90°
I & Q LO
GENERATIO
MSK CARRIER AND TIMING RECOVERY
MSR
MSK CORRELATORS & 90° PHASE RESOLUTION
MSR
FSK DISCRIMINAT
INTEGRATE & DUMP FILTERS
PHASE CONTROL
TCL
DATA DECODERS
TCL
MODE SELECT SWITCH
OUTPUT INTERFACE
TTI
DATA OUT
CLOCK OUT

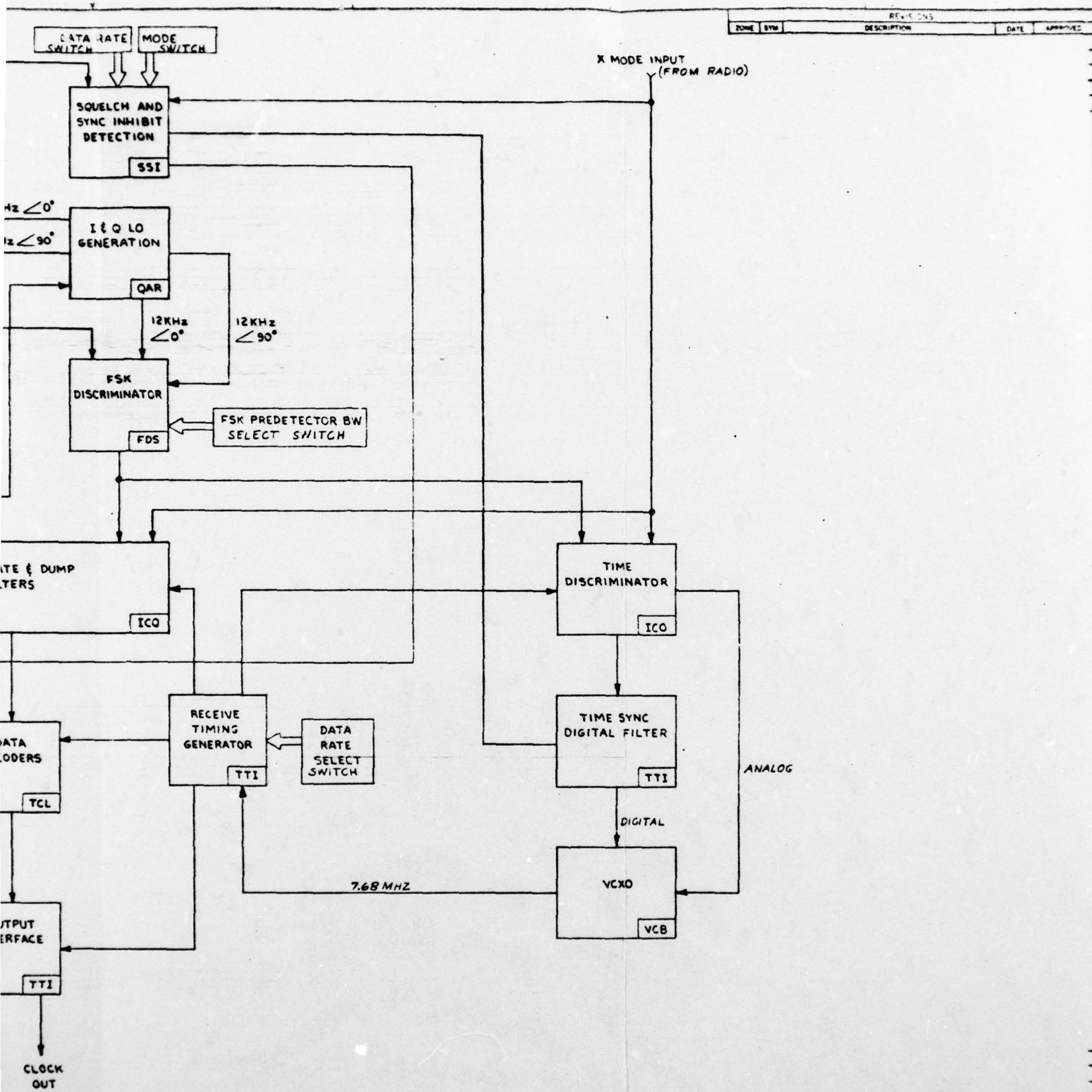


Fig.27 Signal Converter Sys. Block Diagram

Timing Recovery

A time discriminator is used in the timing recovery system. It is used with some modifications in all modes except the MSK mode. The time discriminator is the decision feedback type. It performs well even at very low signal-to-noise ratios.

The loop filter is partially analog and partially digital. The lead path of the filter is analog and the integrator path is digital. Loop bandwidths for the various codes and modulations is shown in Table 5.

Squelch and Sync Inhibit

These two functions are usually performed by measuring a tone outside the signal bandwidth or by "notching out" a piece of the transmitted spectrum so that the noise in this notch can be measured at the receiver. As the baseband converter does not have access to any unused portion of the spectrum a tone measurement scheme is not possible. Similarly, notching out part of the spectrum would introduce a degradation which could lead to erroneous indications of the performance of some signaling schemes. The baseband converter therefore uses a scheme which measures the mean squared error at the decision point as a measure of the signal-to-noise ratio. This type of measurement is more closely connected to error rate than is the measurement of thermal noise and received signal power, since it is valid whether the noise is multiplicative or additive in nature. This technique also introduces no degradations. A front panel meter display, a measure of the mean square error, and a front panel control are used to adjust this to a minimum. This adjustment also adjusts the thresholds used in the multilevel modes. This adjustment is necessary only when the modulation index of the radio is

Table 5

Rate Multiplier Parameters

SYMBOL RATE	DATA RATE Q		DATA RATE X		PLL BW (Hz)		RATE MULT.	
	4LFSK	FSK MSK	PST BIP TWB	MAN DLM CDP	Q	X	Q	X
75	150	75			.5		63/64	
150	300	150			1		32/64	
300	600	300			2		16/64	
600	1200	600			.5		8/64	
1200	2400	1200			1		4/64	
2400	4800	2400	2400		2	8	2/64	32/64
4800			4800	2400		8		16/64
9600			9600	4800		8		8/64
16000			16000			8		8/64
19200				9600		8		4/64
32000				16000		8		4/64

changed or drifts. This display is also useful as a quality monitor.

Mechanical

The eight cards are housed in a card file mounted in a cabinet above the control panel. The power regulators are mounted below the card file and behind the front panel. The units are approximately 14 inches high, 20 inches wide, and 26 inches deep.

3.2 Equipment Description

3.2.1 Quasi Analog Signal Generation

In the Quasi Analog (QA) mode, three types of signals are generated. These are: Frequency Shift Keying (FSK), Minimum Shift Keying (MSK) and Four Level Frequency Shift Keying (4LFSK). The data rates provided are 75, 150, 300, 600, 1200, and 2400 bps for the FSK and MSK and 150, 300, 600, 1200, 2400 and 4800 bps for the 4LFSK.

It should be noted that the notation FSK does not completely define the modulation form unless the modulation index is also specified. This unit provides for modulation indices of from 0.2 to 1.9 where the modulation index is defined as:

$$m = (f_2 - f_1)T$$

where: f_2 = highest frequency (Hz)

f_1 = lowest frequency (Hz)

T = symbol duration (sec.).

Figure 28 is a block diagram of the QA mode signal generator. The same unit is used for all four modulation types. This circuitry is implemented on the QAR card. The output of this unit is at a 2 kHz center frequency.

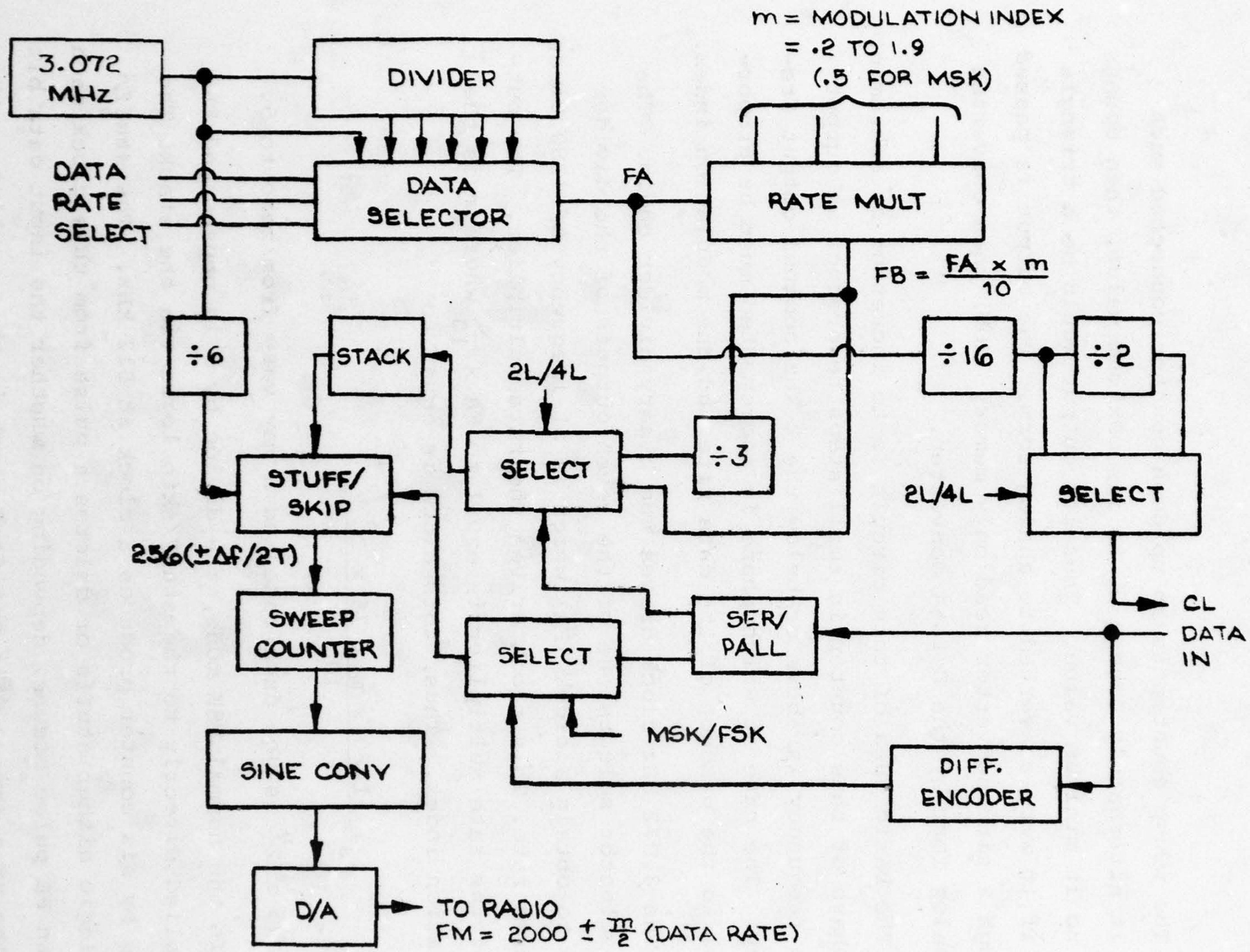


Fig. 28 Block Diagram of QA Mode Signal Generator

The sweep counter is an up/down counter connected such that it alternately counts up to its maximum value, then counts down to it minimum value. Thus its output would be a triangle wave if it were converted to analog form. The output is passed through a sine converter read only memory (ROM) then converted to analog form by the D to A converter.

The basic idea of this circuit is to increment or decrement the phase of this counter in small steps to produce an output whose frequency is above or below the 2 kHz nominal output frequency. The rate at which phase is incremented must be proportional to the product of the data rate and the modulation index.

The 3.072 MHz clock drives the binary divider chain. The data selector selects one of the seven outputs of the divider chain to obtain a clock (FA) which is at frequency of 1280 times the data rate. This clock drives the rate multiplier. The output of the rate multiplier is equal to FA x $\frac{m}{10}$ where m is the modulation index. Thus, this clock is equal to

$$FB = \frac{1280 \times m \times 75 \times 2^n}{10}$$

where 75×2^n is the data rate and n may vary from zero to 5.

In the normal FSK mode, the divide by 3 is unused and FB is applied directly to the stuff/skip logic via the stack. The divide by six counter produces a clock at 512 kHz. The stuff/skip logic either stuffs or deletes a pulse from this clock each time an FB pulse occurs, depending on whether the input data bit is a one or a zero. This modified clock is then applied to the sweep counter. The stack is required since at 2400 bits/sec the FB pulse train has a burst rate which is greater than the

rate at which the stuff/skip logic can accept pulses. The stack accumulates bursts of pulses from the selector and delivers them at a maximum stuff/skip rate. The sweep counter has a period of 1/256 at its input clock; thus it can be seen that each time a stuff or skip pulse occurs, the phase of this clock is altered by $2\pi/256$ radians.

For example, when the modulation index is .5 and the data rate 300 bps,

$$FB = \frac{1280 \times .5 \times 300}{10} = 19200 \text{ Hz.}$$

The number of FB pulses that occur each data interval is

$$N = \frac{19200}{300} = 64.$$

This means that the phase of the sweep counter will be modified by $64/256 \times 2\pi = \pi/2$ radians each data interval, which is correct for this modulation index. If a long sequence of "one's" were transmitted, the output frequency would be

$$F_{out} = \frac{512000 + 19200}{256} = 2075$$

which is 75 Hz above the carrier; again this is correct for this modulation index and data rate.

In the MSK case, the modulation index is always 0.5 and the data is precoded by a differential encoder. Otherwise the operation is the same as the FSK case.

In the four level FSK case, two successive input bits are encoded into one of four symbols. The symbols are 2 kHz + $\Delta/2$, 2 kHz +$\Delta/6$, 2 kHz-$\Delta/6$ and 2 kHz-$\Delta/2$, where $\Delta=f_2-f_1=m/T$. The symbol

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VHF DATA MODEM/BASEBAND DATA SIGNAL CONVERTER.(U)
MAY 78 L EHRMAN, P F MAHONEY, M MICHALIK DAAB07-76-C-0193
UNCLASSIFIED A-216F CORADCOM-76-C-0193 NL

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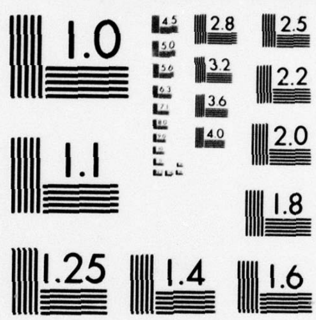


MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS-1963-A

rate is half the bit rate. The first of each two bits determines the sign of the frequency shift and the second bit determines whether it will be $\Delta/2$ or $\Delta/6$.

The $\Delta/6$ frequencies are obtained by dividing the input to the stuff/skip logic by three when these symbols are required. The frequency deviation resulting from this definition of m is one-third that often resulting for 4-LFSK. In particular, "optimum 4LFSK" is often defined as having a modulation index of 0.25; to obtain this case, a setting of m=0.7 or 0.8 should be used, as 0.7/3 and 0.8/3 are the two available indices which are closest to 0.25.

3.2.2 Stable Oscillator and L.O.Generators

The 3.072 MHz input of the QA mode signal generator is derived from a 15.36 MHz stable crystal oscillator. This unit has a stability of better than 1 part in 10^7 per day. The 3.072 MHz is developed by dividing the output of this by 5. The 15.36 MHz oscillator is mounted to a panel at the rear of the cabinet.

Three other signals must be developed from this oscillator and these are: the 128 Δf clock used by the MSK carrier recovery loops, the 10 KHz L.O. for the single sideband up converter and the 12 KHz L.O. for the baseband frequency discriminator. Two clocks, 90^o out of phase, must be generated for each of the local oscillators.

The circuitry to develop these clocks is located on the QAR card. The 128 Δf clock is obtained from the output of the rate multipliers. When the system is being used in the MSK mode the modulation index is forced to 0.5, which causes the rate multiplier output to be the correct frequency.

The 12 kHz clocks are developed by dividing the 3.072 MHz by 256. The last stage of division is performed twice, once with the clock inverted and once without. This produces two clocks 90^{o} out of phase.

The 10 KHz L.O. is developed by dividing the 15.36 MHz by three, then 256, then two. The divide by two is performed twice, once with the clock inverted to produce two 10 KHz tones 90^{o} out of phase.

3.2.3 X-Mode Signal Generation

The X-mode signal generator produces the six codes required in the X-mode: twinned binary,bipolar, paired selected ternary, Manchester, conditioned diphase and delay modulation.

This circuitry is located on the XSG card. The clock generation input is a 192 KHz clock derived from the 15.36 MHz stable crystal oscillator on the QAR card. The 192 KHz is divided by 6, 10, 20 and 40 to produce 32 KHz, 19.2 KHz, 9.6 KHz and 4.8 KHz. These clocks are gated by the data rate select switches to select the clock which is twice the data rate. The selected clock is divided by two and four to produce clocks at the data rate and half the data rate. These clocks are used to generate the various codes and to code the transmit data in via the I/O network on the TTI card.

3.2.4 Single Sideband Upconverter

In the Quasi Analog mode the input signal has a carrier frequency of 2 kHz. To perform proper frequency discrimination on

the FSK signals and matched filtering of the MSK signals at this IF frequency over the required range of data rates and modulation indicates would be difficult if not impossible because of the aliasing problem. Therefore the signal is mixed to a 12 kHz IF by means of a single sideband (SSB) operation. The circuitry which performs this operation consists of a delay equalizer located behind the front panel beneath the card cage, and circuitry located on the FDS card. The circuit is a phase shift type single sideband network. This type of SSB modulator requires a network that produces a 90° relative phase shift at all frequencies in the band of interest. The circuit shown has a theoretical phase shift error of less than one degree from 300 to 4000 Hz.

The envelope delay from input to either output of the 90° relative phase shift network is frequency dependent. This causes asymetric waveform distortion at the frequency discriminator output, and contributes to intersymbol interference. The delay equalizer predistorts the envelope delay of the inputs to the two phase shift networks, thereby linearizing the composite frequency response.

A pair of local oscillators tones shifted inphase by exactly 90° is also required. These are produced by divider networks located on the QAR card. The input to these divider networks is the stable crystal oscillator.

The outputs of the phase shift networks are multiplied by the 10 kHz local oscillator tones, and then summed in an operational amplifier to produce the desired output centered at a carrier frequency of 12 kHz. The unwanted lower sideband, at 8 kHz, is suppressed more than 30 dB by this circuit.

3.2.5 FM Discriminator

The FM discriminator accepts the received FSK signal after it has been shifted to a 12 KHz carrier by the SSB upconverter.

The FM detection is done by means of a baseband discriminator preceded by a front panel selectable low pass filter. This technique is used because of the wide variety of data rates and modulation indices which must be detected. At the lowest data rate and smallest modulation index the unit must detect frequency differences of ± 7.5 Hz; at the highest data rate and modulation index the frequency differences are ± 4560 Hz. At the higher frequencies the operation will be limited by the bandwidth of the radio to lower modulation indices.

To perform accurate FSK detection over this range it is necessary to have a discriminator whose center frequency does not drift by an amount significant relative to the lowest frequency difference to be detected. In the present case this means the discriminator must be accurate to within a small fraction of a Hertz. This can be achieved by means of a baseband discriminator if the L.O.'s are sufficiently accurate. The L.O.'s used are derived from a highly stable oscillator accurate to one part in 10^7. The signal at the transmitter is similarly derived from a highly accurate source. Therefore the L.O.'s are not a source of error. Predetection baseband filtering is performed in each leg of the discriminator. The bandwidth of these low-pass filters is selectable by a front panel thumbwheel switch. These filters remove the double frequency components and adjust the noise bandwidth of the input to match the signal bandwidth. The bandwidth is adjustable from 50 to 3950 Hz in 50 Hz steps.

The outputs of the low pass filters are then differentiated, and the output of each differentiator is multiplied by the input to the differentiator in the other branch. These two products are then summed to form the output. The output of this network is a voltage which is proportional to the frequency difference between the input signal and the 12 kHz local oscillator.

The output of the FM discriminator is integrated in an integrate and dump filter located on the ICO card. The data decision is made on the output of the integrated and dump.

The differentiator must operate over a wide range of frequencies. To ease the requirements on this circuit, a different capacitor is used for data rates below 600 symbols per second by switching in a capacitor in parallel with the one used for the higher symbol dates.

3.2.6 MSK Receiver

Although the transmitted waveform in a MSK system is the same as an FSK system with a modulation index of 0.5, and in fact the same modulator can be used, the MSK receiver structure is very different from the FSK structure. Figure 29 is a block block diagram of the MSK receiver structure. This circuitry is located on several cards. The circuitry unique to the MSK receiver is located on the MSR card.

The same SSB converter that is used for the FSK detection is used to convert the input signal to a 12 KHz carrier frequency. The output of the SSB converter is passed through a squarer to produce a waveform which has a specular components at $2F_1$ and $2F_2$ where:

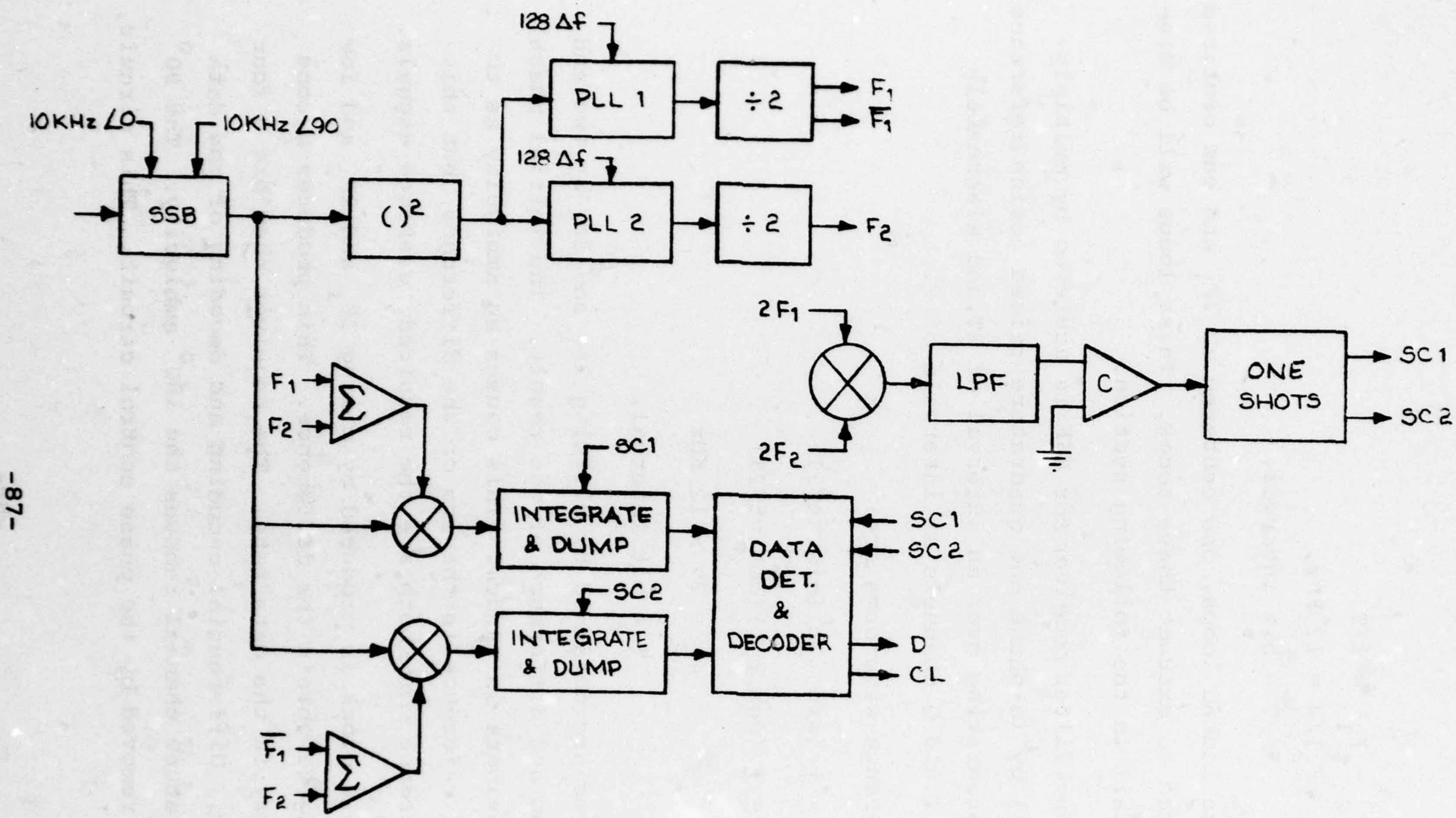


Fig. 29. MSK Receiver

$$2(F_2 - F_1) = 1/T,$$
$$(F_1 + F_2)/2 = 12 \text{ KHz},$$
and $$T = \text{bit interval.}$$

Two phase locked loops, one centered at $2F_1$ and one centered at $2F_2$ are used to extract these tones. These loops will be discussed in detail in the following section.

The matched filter receiver for MSK is achieved by multiplying the signal by in-phase and quadrature raised cosine reference waveforms, integrating over an interval of 2T, and alternately sampling the I and Q channels at intervals of T.

The reference waveforms are:

$$IR = \pm (\sin \omega_c t)(\sin \pi t/2T)$$
$$QR = \pm (\cos \omega_c t)(\cos \pi t/2T)$$
$$\omega_c = 2\pi \times 12 \text{ KHz}$$
$$T = \text{bit interval.}$$

These waveforms are obtained by dividing $2F_1$ and $2F_2$ by two and taking the sum and difference of the result. The initial phase of the two dividers is random. This causes an ambiguity as to whether the I reference is the sum or the difference, but this makes no difference since both are the required reference signals.

The sample clock is produced by mixing $2F_1$ and $2F_2$ and low pass filtering to obtain the difference. This produces a tone whose frequency is the data rate. The receiver then has a four way ambiguity. Differential encoding and decoding of the data on each quadrature channel removes the 180° ambiguity. The 90° ambiguity is removed by the phase control circuit. This circuit

examines the reference waveform on the I channel. If the nulls of this waveform are coincident with the sampling interval it does nothing because this indicates the correct timing. If the nulls are not coincident with the sampling time the clock is shifted to make it so.

The signal is mixed with the two reference waveforms and then integrated by integrate and dump filters located on the ICO card. The output of the integrate and dump is sampled and a decision made then. The data is differentially decoded and converted to a single bit stream by a parallel to serial converter. The decoding and parallel to serial conversion is performed on the TCL card.

3.2.7 MSK Carrier Recovery Loops

The carrier recovery loops in the MSK receiver must have very narrow bandwidths and must operate at a different frequency for each data rate. Normally narrow phase locked loops are designed using VCXO's. This would require a different VCXO for each data rate, a total of twelve, which would be cumbersome and expensive. To avoid this, a pair of tones, one at 128 Δf and one at 128 x 24 kHz, are extracted from the dividers on the Quasi Analog signal generators and used to make a source whose frequency may be varied just slightly around the desired frequency.

3.2.8 Integrator Combiner

The Integrator Combiner card is used in all modes, both QA and X. In all modes a signal must be integrated over an interval then compared to one or more than one threshold to make the data decision. The ICO also contains the time discriminator circuitry

which is used for timing recovery in all modes except the MSK mode.

3.2.9 Data Decoders

The data decoders for the Manchester, Conditional Diphase and Delay Modulation are located on the ICO card. The FSK mode requires no decoding after the digitization, so it is also taken from the ICO card. The other modes are decoded on the TCL card.

3.2.10 Time Discriminator

The time discriminator is located on the ICO card. Figure 30 is a block diagram of the discriminator. As shown in the block diagram the time discriminator shares the two upper integrators with the data detections. The time discriminator operates as follows. Alternate input symbols are integrated by I1 and I2. At the end of each symbol, a decision is made as to whether or not a transition occurred. The lower two integration is switched into the sample and hold amplifier. If a transition is detected the sample and hold reads the signal. If a positive going transition is sensed then the output switch selects the output of the sample and hold, while if a negative going transsition is sensed, the output of the -1 amplifier is switched to the output.

It can be seen that when the timing loop is perfectly locked the outputs of I3 and I4 will integrate to zero whenever a transition occurs. If a small phase error occurs the integrator outputs will be nonzero by an amount proportional to the phase error. The direction of the transition will determine the sign of the integrator output. This effect is removed by the output switch.

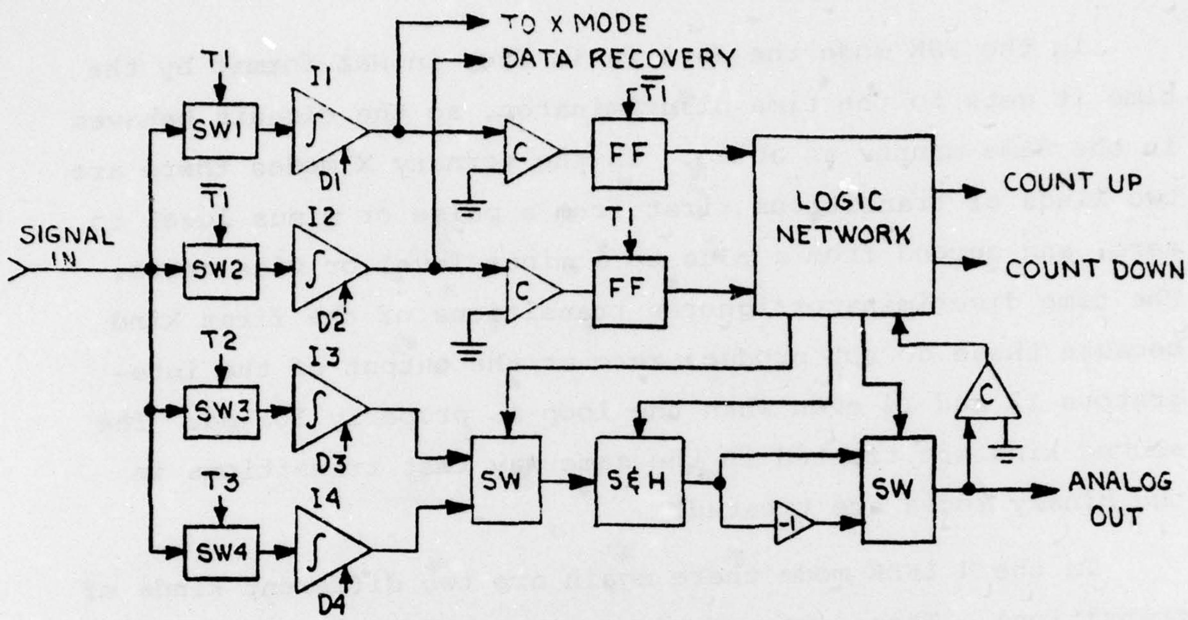


Fig. 30 Timing Discriminator

The loop filter requires two inputs, one an analog voltage proportional to the phase error and another which is simply the sign of the phase error.

The sign of the phase error is developed by the LM 311 Comparator at the output.

The logic circuitry which produce the sample and switch pulses are located on the TCL card.

The phase discriminator operates slightly differently for different modes. In the binary modes it acts exactly as described above. It essentially assumes that the input data is NRZ and locks onto the transitions. Further processing, to resolve the phase, is necessary because the transitions occur at twice the bit rate. This is done on the TCL card.

In the FSK mode the data is in fact in NRZ format by the time it gets to the time discriminator, so the circuit behaves in the same manner as above. In the ternary X modes there are two kinds of transitions first from a pulse or minus level to zero, and second from a plus to a minus level or vice versa. The time discriminator ignores transitions of the first kind because these do not produce zero at the output of the integrators I3 and I4 even when the loop is properly locked. The second kind are treated in the same way that transitions in the binary modes are treated.

In the 4 LFSK mode there again are two different kinds of transitions. These are non-symmetrical and symmetrical transitions. Non-symmetrical transitions are ignored by the time discriminator and symmetrical transitions are treated the same as transitions in the binary modes.

In the MSK mode the time discriminator is not used because the timing is derived from the carrier recovery loops.

3.2.11 Time Synchronization Loop

The Time Synchronization Loop recovers the timing for all modes except the MSK mode.

The loop consists of the time discriminator, the time sync digital filter, the VCXO box, and the receive timing generator.

The time discriminator circuitry is located on the ICO card and is described in Section 3.2.10. The time sync digital filter is located on the TTI card as is the receive timing generator. The VCXO box is located behind the card file.

Figure 31 is a block diagram of the time synchronization system.

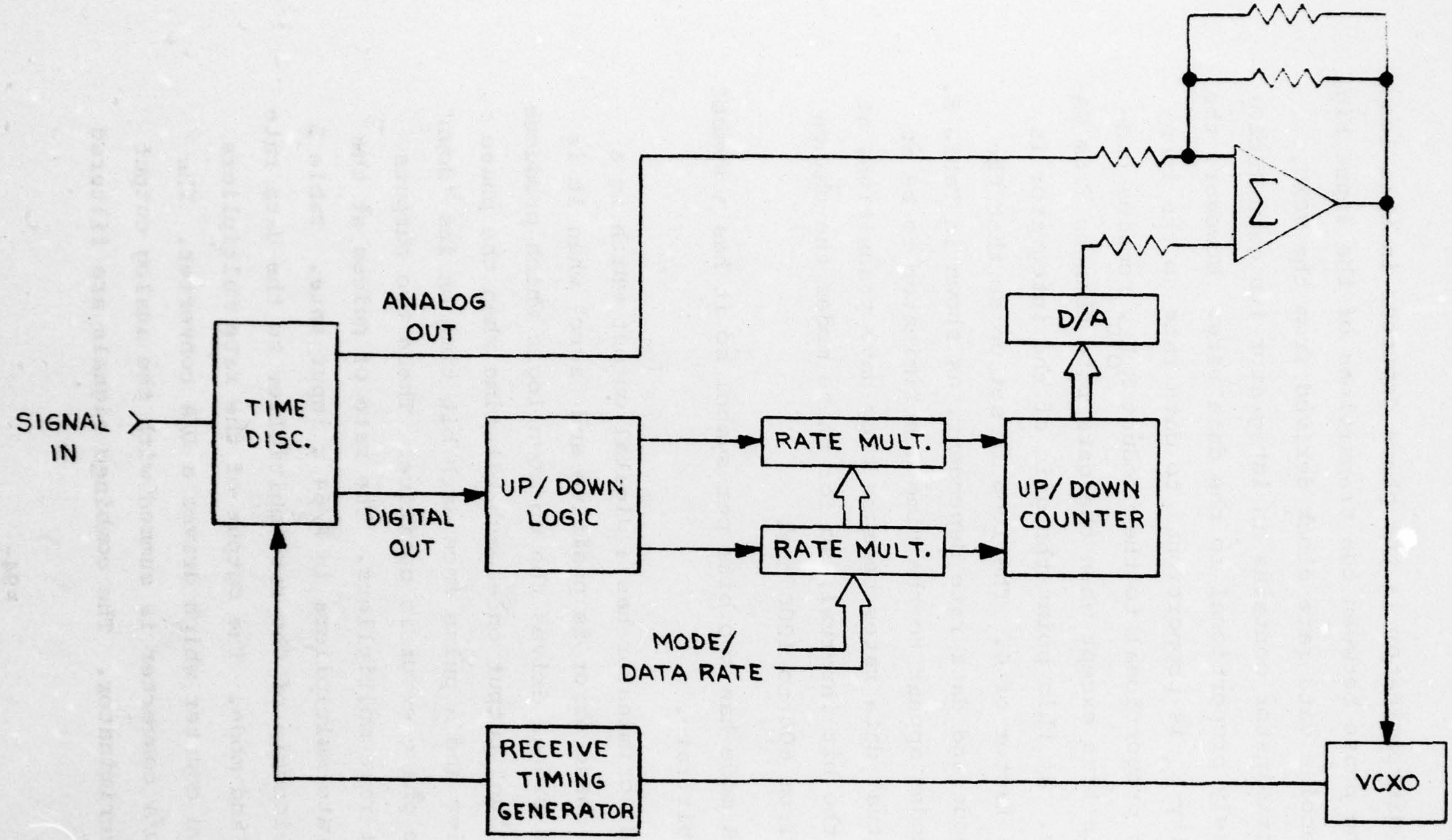


Fig. 31 Time Synchronization System

The time discriminator is the phase detector in this loop. It measures the phase between the transitions of the input signal and the receive data rate clock derived from the VCXO. Since the discriminator contains an integrator its gain factor (K_D) is inversely proportional to the data rate. However, the VCXO sensitivity K_O is proportional to data rate so the loop gain, which is proportional to the product $K_O K_D$, remains constant. This is true except when the data rate changes from QA-mode to X-mode. At this point the gain of the integrator is increased by a factor of 4. The loop is set up so that the bandwidth is mode and data rate dependent, as shown in Table 5. The binary X-modes appear to the time discriminator to be at twice their actual data rates because they have transitions at the center of the bit interval, so for these modes the change is made going from 600 to 1200 bps.

The 4 LFSK mode has two bits per symbol so it has a symbol rate half its bit rate.

The time discriminator has a digital output which is a "one" when the phase error is positive and "zero" when it is negative. This signal drives the up/down logic which produces a pulse on its "up" output once each bit time when the phase error is negative and a pulse once each bit time on its "down" output when the phase error is positive. These two outputs drive a pair of rate multipliers. The rate of pulses at the output of the rate multipliers is R/64 x input rate. Table 5 shows the relationship of the rate multiplier to the data rate loop bandwidth and mode. The output of the rate multipliers drive an up down counter which drives a D/A converter. The output of the D/A converter is summed with the analog output of the time discriminator. The combined signals are filtered

to remove any high frequency "glitches". The bandwidth of this filter is made greater than the loop bandwidth so it does not affect the loop dynamics. The output of the filter drives the VCXO. The output of the VCXO drives the receive clock generator, which is simply a divider network which is controlled by the data rate and mode select switches. The output of the receiver clock generator drives logic on the TCL card which produces timing for the time discriminator.

The VCXO operates at an actual center frequency of 7.68 MHz. It has an input voltage range of ± 5V and an output frequency swing of ± 770 Hz. Thus the sensitivity is

$$K_O = 154 \times 2\pi \text{ rad/sec/volt.}$$

The output of the VCXO is divided down to the data rate so it has an equivalent sensitivity of

$$K_O' = \frac{154 \times DR}{7.68 \times 10^6} \times 2\pi \text{ rad/sec/volt}$$

where

$$DR = \text{symbol rate} = \text{data rate.}$$

The time discriminator has a sensitivity of

$$K_D = \frac{12000}{2\pi\ DR}\ \frac{\text{volts}}{\text{rad}}$$

for the higher data rates. Therefore

$$K_O' K_D = \frac{154 \times 12000}{7.68 \times 10^6} \text{ sec}^{-1}$$

$$= 0.24 \text{ sec}^{-1}.$$

For a first order loop

$$H(s) = \frac{K}{s+K}$$

$$K = K_D K_O' \times K_f = \text{radian cutoff frequency}$$

for

$$\omega_{3dB} = 16\pi \text{ rad/sec}, \quad f_{3dB} = 8 \text{ Hz}$$

and

$$K_f = \frac{2\pi\ 8}{K_O' K_D} = \frac{2\pi\ 8}{.24} = 209.$$

The gain K_f of the low pass filter is actually 240. The digital accumulator is made such that it moves slowly relative to the analog path to the VCXO. This is done because of the one bit quantization at the input to the accumulator. Thus the loop acts like a highly overdamped second order loop. A highly overdamped second order loop acts almost exactly like a first order loop with one important exception, the mean phase error is driven to zero. This is the sole function of the digital accumulator. Without this feature it would not be possible to "freeze" the loop when the SNR is less than 6 dB as is required by the specification. When the freeze command is actuated by the SSI card the accumulator content are used to set the VCXO, since the up/down inputs to the accumulator are inhibited. If the loop has been locked for a significant amount of time the contents of the accumulator will be an accurate estimate of the frequency shift of the input signal. Thus the system will remain in synchronization even during long deep signal fades.

3.2.12 Phase Control

If no precautions were taken, it would be possible for the timing system to lock up 1/2 symbol away from the proper timing phase for Paired Selected Ternary (PST), Manchester (MAN), Delay Modulation (DM), Conditional Diphase (CDP), and MSK. Phase control circuits are used to prevent this lockup. The MSK phase control is discussed in Section 3.2.6. The QA mode phase controls operate as follows.

PST has three input data sequences which indicate that the timing phase is wrong. The data occurs in two bit frames when the timing is off, one bit from each of two adjacent frames is decoded. When the timing is correct the improper bit sequences never occur unless a detection error is made. The phase resolution circuit looks at the data both at the instant the data is decoded and 1/2 a frame later. If the timing is correct and no errors are made the "impossible" sequences will never occur at the instant the data is decoded but will occur occasionally 1/2 a frame later. When the timing is off these sequences will never occur 1/2 frame late but will occur some of the time at the time the data is decoded. The Phase Resolution circuit therefore counts the number of times these sequences occur at the correct timing phase. If this counter reaches eight before one of these sequences occurs at the wrong timing phase, the timing is shifted 1/2 a frame.

The Manchester and Conditioned Diphase cases use the same phase control circuit. Both of these codes always have a transition in the center of the baud when the timing is correct. When the timing is incorrect there may or may not be a transition at the center. The Delay Modulation data decoder is operational when either Manchester or Conditioned Diphase

is being received. In Delay Modulation a one is decoded whenever there is a transition in the center of the baud and a zero is decoded when there is no transition. Therefore by examining the Delay Modulation data, one can determine if the timing phase is correct. If the timing is correct, it will put out all ones, while if the timing is wrong, it will put out both zeros and ones.

Therefore, for Manchester or Conditioned Diphase the circuit operates as follows: when the SQUELCH level goes high indicating a new message is being received, the circuit counts for 32 clock pulses. If at the end of this period no zeros have been detected by the Delay Modulation decoder, the system is reset and the clock is left alone. If a zero has been detected the timing is shifted 180° and the process is repeated. This continues until 32 ones in a row are received by the Delay Modulation decoder.

The phase control for Delay Modulation relies on the fact that in Delay Modulation the sequence 101 is never detected when the timing is off by 180° and there are no detection errors. As before the circuit is reset by the squelch line. The clock is counted, and if 32 bits are received without a 101 sequence, it is assumed that the phase of the clock is wrong; the clock is then shifted 180° and another 32 clock pulses are counted. This continues until the 101 sequence is found. The cycle is terminated and the phase is left unchanged until the system is reinitiated by the squelch going off and on. If the squelch is actuated because of a channel fade the cycle will be reinitiated. In this case, however, it will always start out with the correct phase (unless the fade is so long that the time synch loop losses a bit). The probability of the

unit failing the test when the timing is correct is about one in twenty (assuming no errors). Therefore one out of twenty times that the channel fades below the squelch level, the system will lose the second set of thirty two bits after the fade.

3.2.13 Input Output Interface

The input/output circuitry is located on the TTI card.

There are two digital input lines, the Request To Send (RTS) and the NRZ Data In. These two signals are in MIL 188C format. The LM311 comparators accept data in this format. As a convenience an internally adjustable threshold is provided, which allows the threshold to be adjusted so that TTL, CMOS, ECL or any other commonly used logic levels may be used as input signals. For use with MIL 188C type inputs the threshold is set at or near zero. The input data may be inverted by means of a front panel toggle switch, in order to accommodate situations where a phase inversion occurs between transmitter and receiver. The input clock may also be inverted. This is so that the unit can accommodate data sources that trigger on either edge of the clock.

The Push to Talk relay PTT is actuated one clock pulse after a RTS level is received. There is a 150 millisecond delay between the time that the relay is actuated and the Data Enable line goes high. The gated transmit clock pulses are also delayed for 150 milliseconds.

The digital output drivers are 9616 integrated circuit line drivers. These are MIL 188C compatible drivers.

The digital outputs for the transmitter are the Data Enable line, the Gated transmitter timing and the Ungated transmitter timing.

The digital outputs for the receiver are the NRZ output, the gated receive timing and the ungated receive timing.

The gated and ungated received timing may be inverted to allow operation with data receivers which operates on either edge of the clock.

As shown on the drawing the Received Data is selected from the appropriate decoder by gating the decoder outputs with the outputs of the mode select switches. The gated receive timing is gated by the Send level and the Squelch. When the receiver is squelched or when data is being transmitted, this clock is inhibited.

3.2.14 Squelch and Sync Inhibit

The squelch and sync inhibit functions are performed on the SSI card. This circuit computes a measure of the mean squared error at the detection point and compares this to a threshold which is equal to the value of the mean squared error when the signal-to-noise ratio is +6 dB. The result of the comparison is used to inhibit the time synchronization system when the SNR falls below 6 dB. A second comparison is made with a variable threshold to obtain a squelch signal.

The concept of this circuit is as follows. In all modes, the signal converter makes a data decision based on the output of the integrators at a certain instant; call this the detected signal. When no noise is present the value of the detected signal is one of 2, 3, or 4 voltages depending on the mode.

In the presence of noise the decoder determines which level the detected signal is closest to and produces the data bit or bits that correspond to this level. If the detected signal is sampled and held until the data decision is made, the difference between the reference levels and the detected signal can be calculated. In an error free system this would be a perfect measure of the noise. In the presence of data errors the rms value of this signal is monotonic with the signal-to-noise ratio. This signal is called the mean squared error (MSE). Thus it can be seen that this signal can be used as a measure of the SNR which is valid regardless of whether the noise is multiplicative or additive and which requires no degradation of the data detection process.

The reference levels of the detected signal are different for each data rate and modulation mode, and in the FSK modes they are dependent on the modulation index. A means must be provided to generate these levels for each combination. Threshold levels for the ternary and 4 LFSK modes must also be generated. Figure 32 shows the reference level generator. The reference levels for the QA modes are generated by the resistor matrix on the right and those for the X modes by the matrix on the left. For each data rate and mode a resistor is selected which sets the reference and threshold levels for this combination. The D to A converter is controlled by the modulation index so that the voltage across the selected resistor is proportional to the modulation index. The X mode reference levels are produced in a similar way, except that the DAC is not present. The switch in the center is controlled by the X/QA select switch. The pot at the output of this switch is a front panel adjustment that allows the operator to trim the

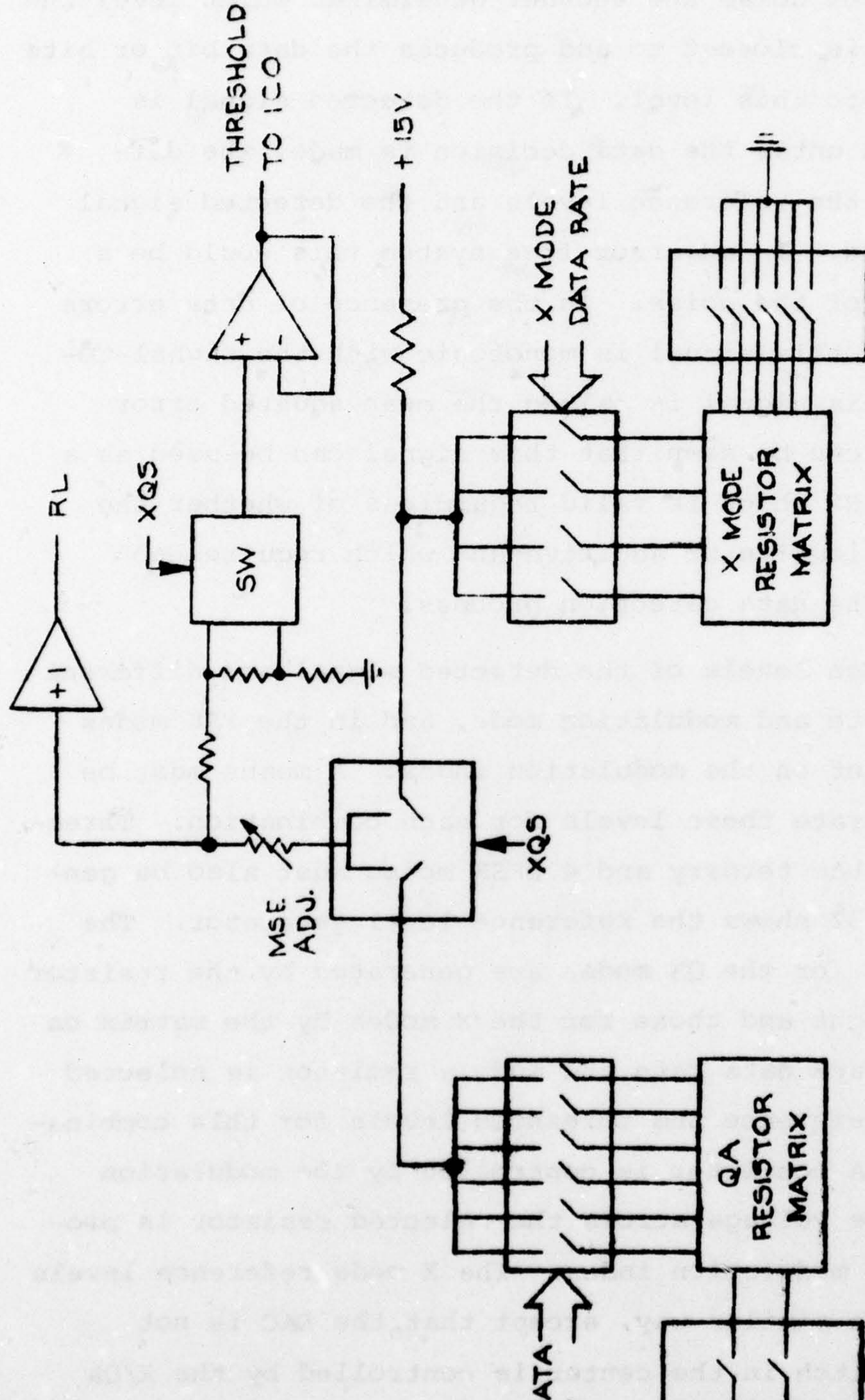


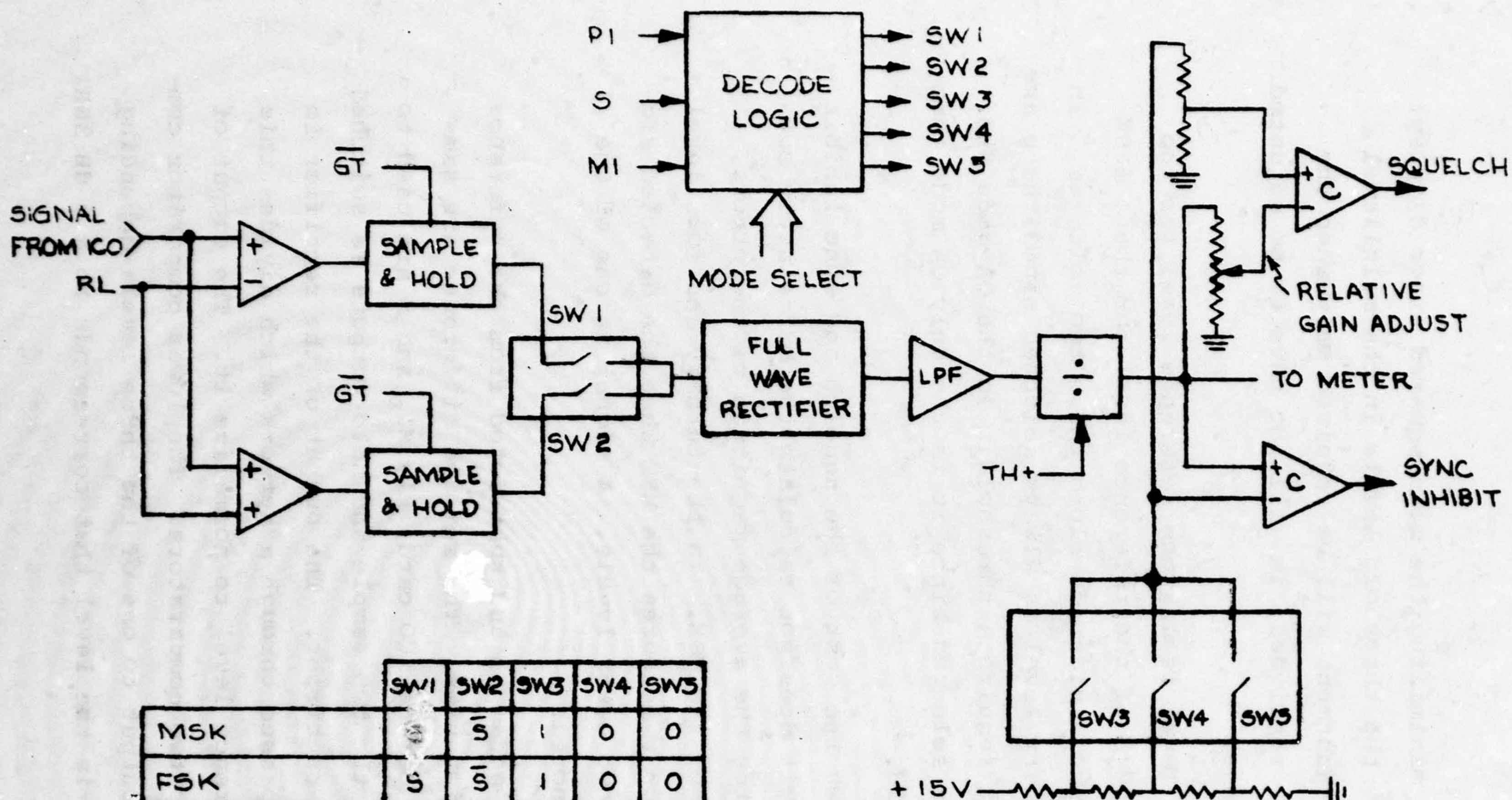
Fig. 32. Reference Level Threshold Generator

reference level by minimizing the mean squared error display. This also optimizes the threshold levels in the multilevel modes. The trim adjustment will be required only when the radio is changed or realigned; it will not have to be adjusted for every mode change.

The threshold level is 1/2 the reference level for the ternary modes, and 2/3 of the reference level for the 4 LFSK mode. The resistance divider develops these two voltages. In the X mode the ternary level is always selected since these are the only modes that require a threshold. In the QA mode the 4 LFSK threshold is selected since it is the only QA mode that requires a threshold.

Figure 33 shows the rest of the squelch and sync inhibit circuit. The circuit does not calculate the MSE exactly but instead it calculates the average magnitude of the errors, which is a measure of the MSE. In the ternary and four level modes the circuit only measures the MSE when the detected signal is one of the two outer levels. A signal at one of the inner levels is ignored.

The signal is added to and subtracted from the reference level at the decision time. The sum and difference are sampled. The outputs of the ICO card, P1, M1, and S are used to determine which of the two sample and hold outputs is switched into the full wave rectifier. The output of the rectifier is filtered and then passed through a network which divides this signal by the reference level to normalize it. The output of the divider goes to two comparators. The lower comparator compares the divider output to one of the three levels depending on the mode. This is the level that corresponds to a 6 dB SNR.



	SW1	SW2	SW3	SW4	SW5
MSK	S	$\overline{S}$	1	0	0
FSK	S	$\overline{S}$	1	0	0
4 LFSK	P1	M1	1	0	0
BILEVEL X	S	$\overline{S}$	0	1	0
TRILEVEL X	P1	M1	0	0	1

Fig. 33. Squelch and Sync Inhibit Generator

It is the same for all modes that have the same number of levels; thus there are three values, one each for the binary, ternary, and quarternary modes.

The second comparator compares the same two signals but has an adjustment which allows the relative gain to be adjusted. This is the squelch control knob on the front panel. The meter is driven by the output of the divider. The meter is used to adjust the gain of the reference level generator when the radio is changed or when the deviation of the ratio is altered. It also serves as a quality monitor.

3.2.15 Mechanical

The signal converters are supplied in stand-alone table top cabinets. The assemblies within the cabinet are a control panel, a card file, and the power regulator assembly. The cabinet is approximately 14 inches high.

3.2.16 Power Supplies

The signal converter power supply is shown on Figure 34. All circuitry within the signal converter operates from ±15 volts or +5 volts.

The input 117V nominal voltage is fused, then passed through an RFI filter. The output of the filter drives three power supplies. One of the supplies has 5 volt output, the others have 15 volt outputs. All outputs are short circuit protected and an overvoltage crowbar is supplied at each output.

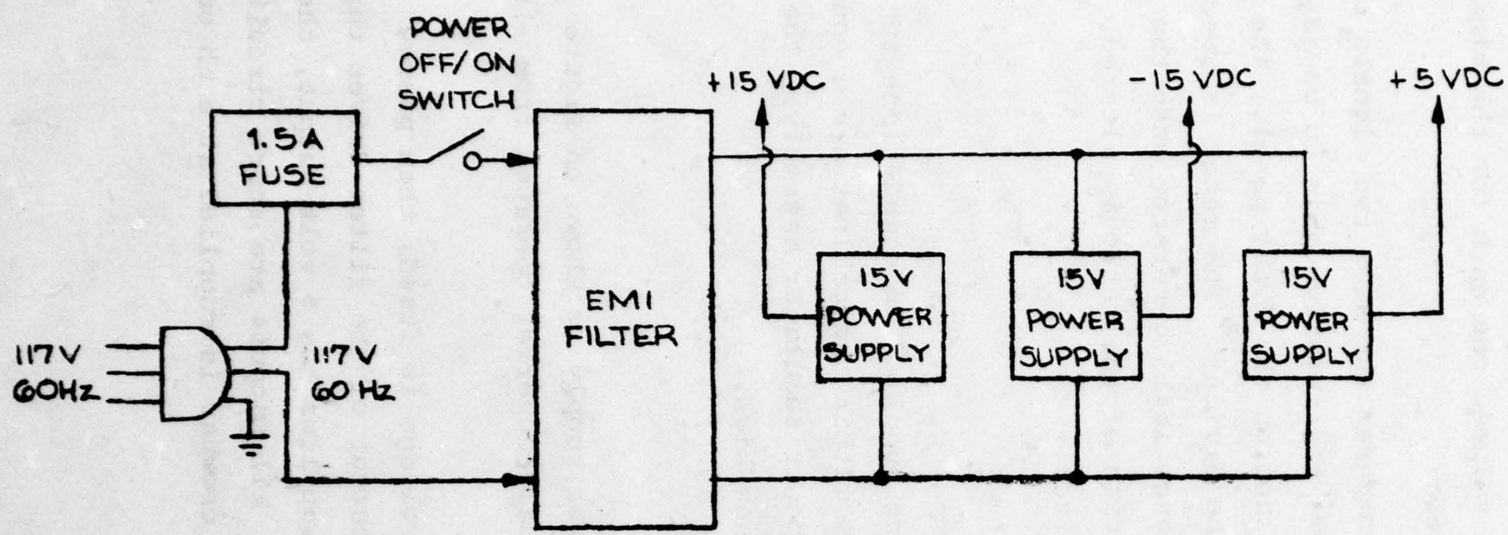


Fig. 34. Signal Converter Power Supply

SECTION 4

TEST RESULTS

4.0 Introduction

In this section we present the results of two types of measurements performed on the Baseband Converters. Baseband and RF spectra are shown in Section 4.1, and error rates in Section 4.2.

4.1 Spectrum Measurements

4.1.1 Baseband Spectra

Sections 2.2 and 2.3 showed computed spectra for quasi-analog and X-mode signals. Spectrum measurements were made of the converter output signals, using a Hewlett-Packard Model 141T Spectrum Analyzer. Table 6 lists the baseband spectra and the spectrum analyzer horizontal scale. All photographs were made at 10 dB/vertical division.

As shown in Section 2, pair-selected ternary, bipolar, and twinned-binary all have their first spectral minima at a frequency equal to the bit rate. Figures 35 through 37 show the spectra for these waveforms, at bit rates of 9600 bps and a spectrum analyzer setting of 2 kHz/div. Thus, the first minima should be and are 4.8 divisions away from DC. Similarly, the spectra of Manchester and Conditioned diphase should have their first minima at a frequency equal to twice the bit rate. Figures 38 and 39 show these spectra, at bit rates of 9600 bps and 5 kHz/div. Thus the first minima should be and are 3.8 divisions away from DC.

TABLE 6

BASEBAND SPECTRA MEASUREMENT PARAMETERS

Figure	Code	Bit Rate	Mod. Index	kHz/Div.
35	PST	9600	–	2
36	BIP	9600	–	2
37	TWB	9600	–	2
38	MAN	9600	–	5
39	CDB	9600	–	5
40	DLM	9600	–	5
41	2-LFSK	1200	1.0	0.5
42	2-LFSK	1200	0.5	0.5
43	4-LFSK	2400	0.8(≡0.27)	0.5
44	4-LFSK	2400	1.9(≡0.63)	0.5
45	MSK	1200	0.5	0.5
46	MSK	2400	0.5	0.5

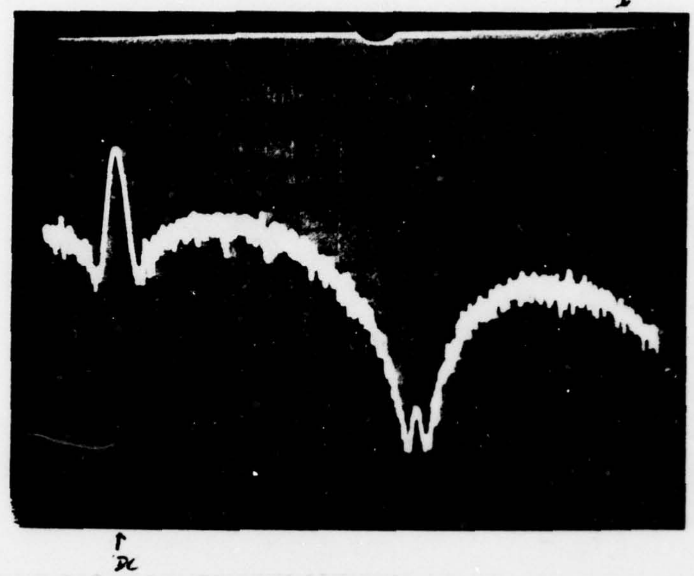


Fig. 35. Pair Selected Ternary Spectrum
9600 bps, 10 dB/Div. Vert.,
2 kHz/Div. Horiz.

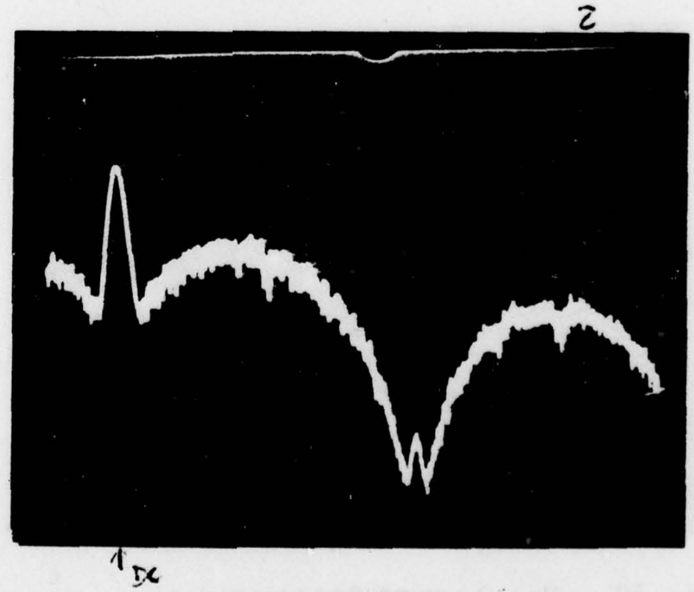
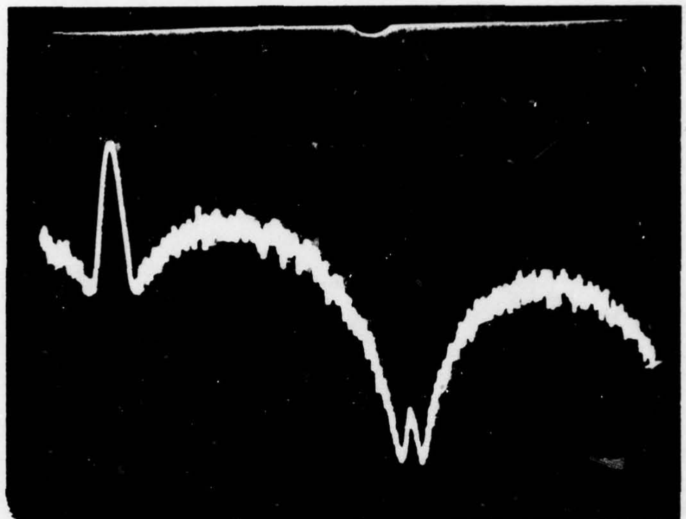


Fig. 36. Bipolar Spectrum
9600 bps, 10 dB/Div. Vert.,
2 kHz/Div. Horiz.

3



fc

Fig. 37. Twinned Binary Spectrum
9600 bps, 10 dB/Div. Vert.,
2 kHz/Div. Horiz.

4

fc

Fig. 38. Manchester Spectrum
9600 bps, 10 dB/Div.,
5 kHz/Div. Horiz

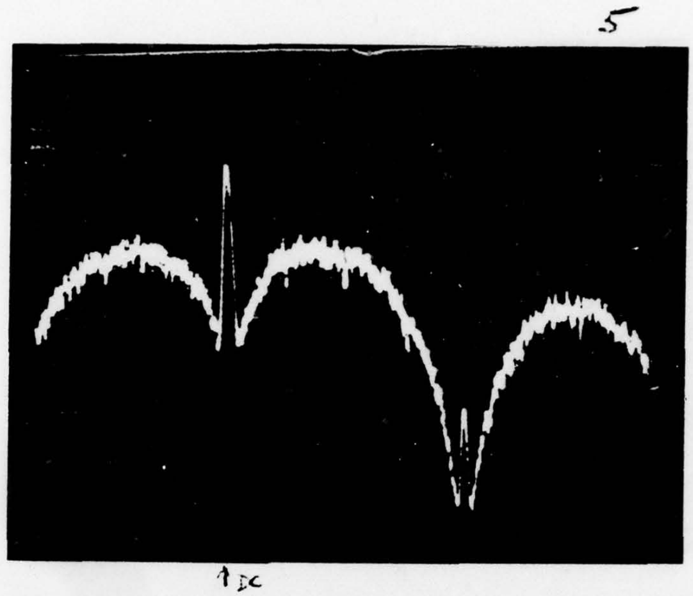


Fig. 39. Conditioned Diphase Spectrum
9600 bps, 10 dB/Div. Vert.,
5 kHz/Div. Horiz.

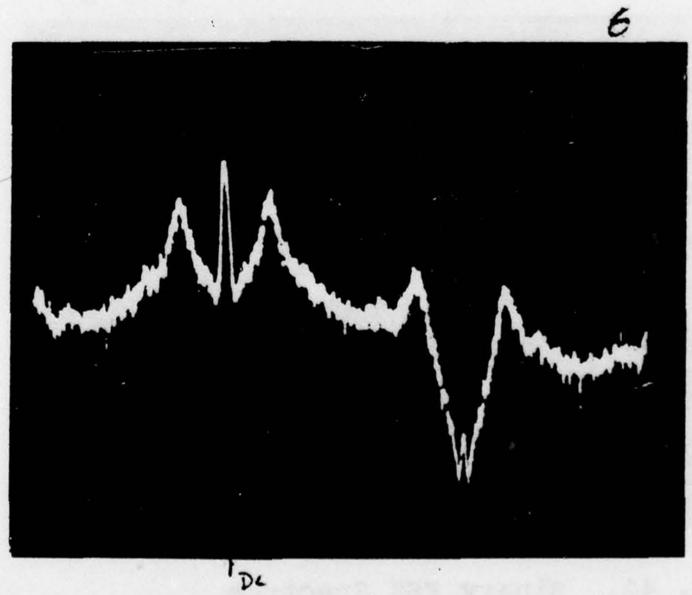


Fig. 40. Delay Modulation Spectrum
9600 bps, 10 dB/Div. Vert.,
5 kHz/Div. Horiz.

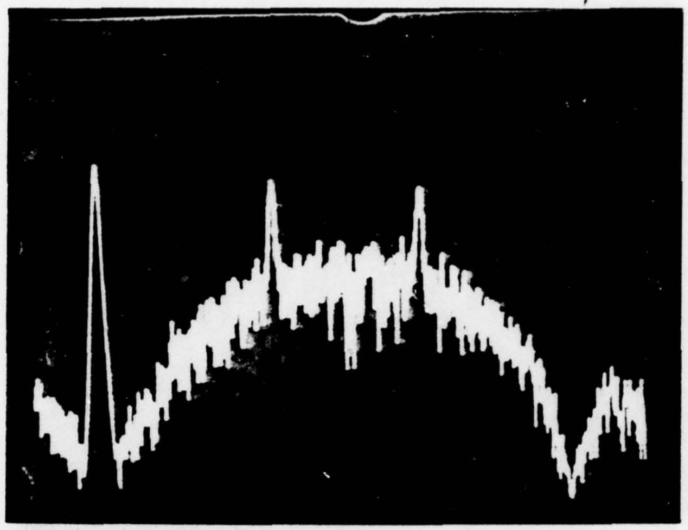


Fig. 41. Binary FSK Spectrum
1200 bps, Mod. Index = 1.0,
10 dB/Div. Vert., 500 Hz/Div. Horiz.

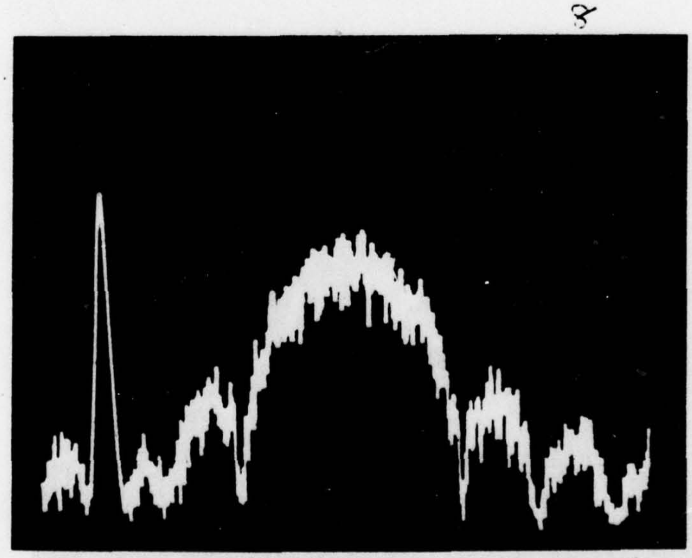


Fig. 42. Binary FSK Spectrum
1200 bps, Mod. Index = 0.5,
10 dB/Div. Vert., 500 Hz/Div. Horiz.

9

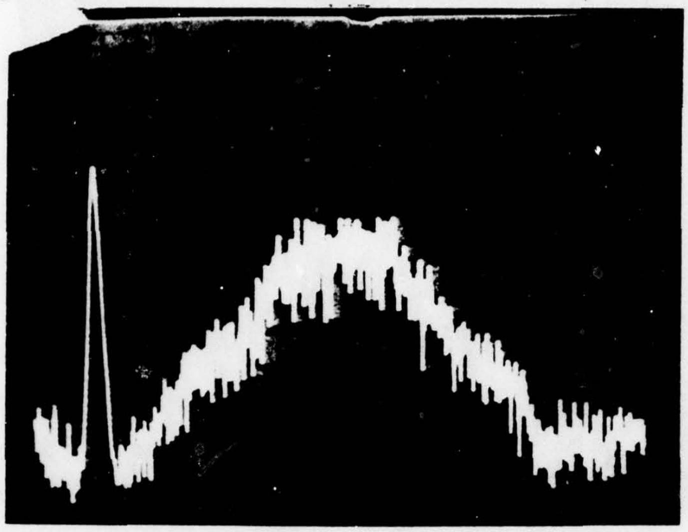


Fig. 43. 4 Level FSK Spectrum
2400 bps, Mod. Index = 0.27,
10 dB/Div. Vert., 500 Hz/Div. Horiz.

10

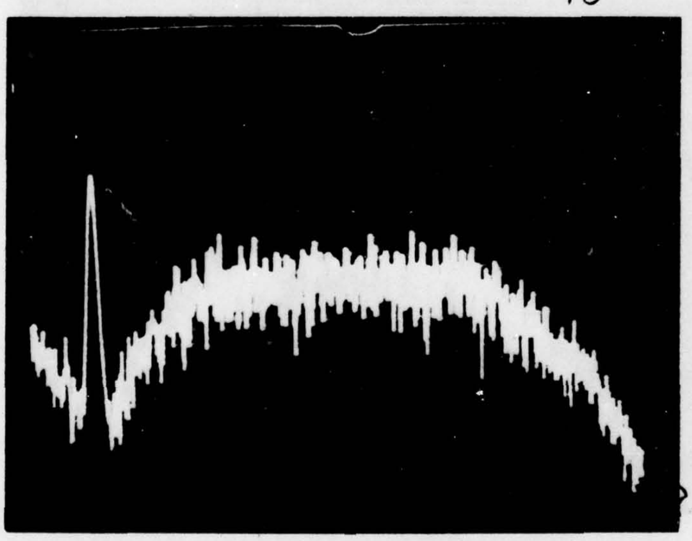


Fig. 44. 4 Level FSK Spectrum
2400 bps, Mod. Index = 0.63,
10 dB/Div. Vert., 500 Hz/Div. Horiz.

11

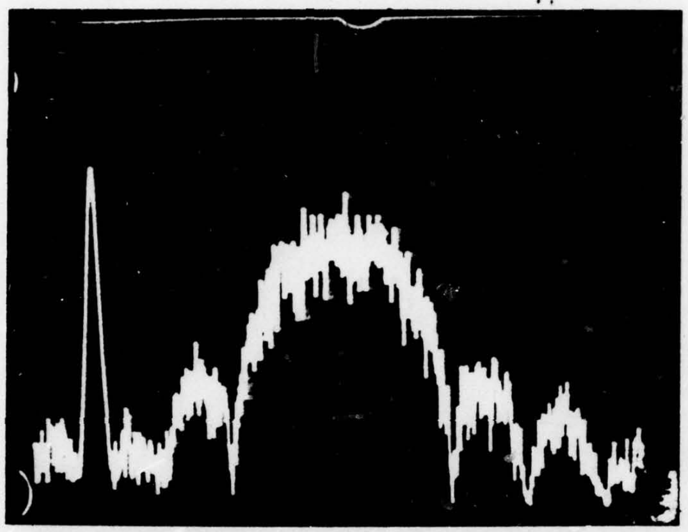


Fig. 45. MSK Spectrum
1200 bps, 10 dB/Div. Vert.,
500 Hz/Div. Horiz.

12

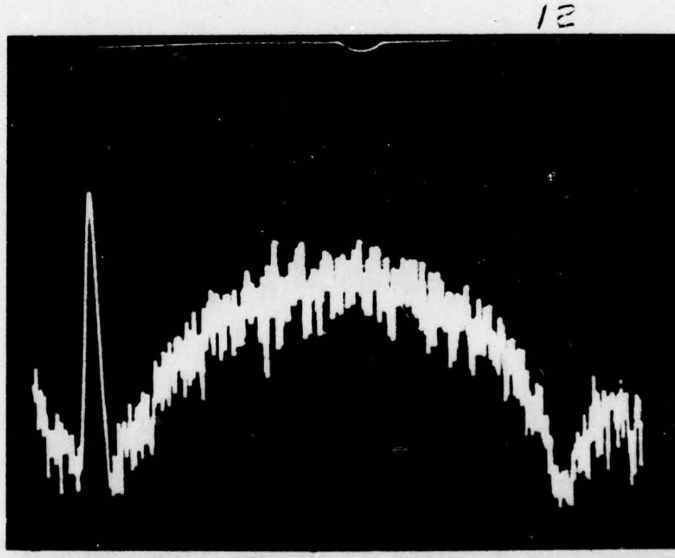


Fig. 46. MSK Spectrum
2400 bps, 10 dB/Div. Vert.,
500 Hz/Div. Horiz.

The spectra of pair-selected ternary, bipolar, twinned binary, Manchester, and conditioned diphase are all similar in shape, aside from a scaling factor. Delay modulation, however, is quite different, having a minimum at DC, and a first maximum at approximately 0.4 times the bit rate, and its first null at twice the bit rate. The measured spectrum of delay modulation is shown in Figure 40, at a bit rate of 9600 bps and a scale of 5 kHz/div. Thus the first maximum should be, and is, 0.8 divisions away from DC, and the first null should be and is 3.8 divisions away from DC.

The X-mode spectra are shown in Figures 41 through 46. Binary FSK are shown in Figure 41, for a modulation index of 1.0, and in Figure 42 for a modulation index of 0.5. Figure 41 shows line spectra spaced away from the 2 kHz FSK center frequency by half the bit rate, or 600 Hz. This is characteristic of coherent FSK with an index of 1. When the index is 0.5, as shown in Figure 42, the spectrum is much more compact with the first null appearing 0.75 times the bit rate away from the center frequency. The 4 LFSK spectra are shown in Figures 43 and 44, for two indices, 0.8/3 and 1.9/3, respectively. As was shown in Section 2, 4 LFSK, with a modulation index of 0.25, has a spectrum which is similar to that of binary FSK with a modulation index of 0.5 and half the data rate. Thus, the spectra of Figures 42 (binary FSK, 1200 bps mod. index = 0.5), and Figure 43 (4 LFSK, 2400 bps, mod. index = 0.27) should be similar. Comparing the two, it is seen that they are essentially the same to the -20 dB points, and that an envelope connecting the sidelobe maxima of the binary FSK matches the 4 LFSK even further. Figure 44 shows the effect of increasing the 4 LFSK modulation index even further; at 2400 bps and an

index of 0.63, the 3 dB spectrum width is in the order of 3 kHz, and would be unsuitable for use in a voice bandwidth channel.

Figures 45 and 46 show MSK spectra for 1200 and 2400 bps. The 1200 bps spectrum is the same as that of Figure 42, binary FSK with an index of 0.5. The 2400 bps spectrum of Figure 46 should be compared to the 4 LFSK 2400 bps spectrum of Figure 43; it is seen that the latter spectrum is more compact, rolling off more quickly away from the 2 kHz center frequency.

4.1.2 RF Spectra

The RF spectra of quasi-analog and X-mode signals, after frequency modulation in the transmitter, are shown in Figures 47 through 54 for the parameters shown in Table 7. The binary FSK and 4 LFSK are shown in Figures 47 through 49, all at 1200 bps. The theoretical spectra of Section 2.4.2 show that the width of the RF spectra is relatively independent of the quasi-analog signal data rate and modulation index; their effects show up primarily in the fine structure of the spectra. The binary FSK FM appears to have more detailed fine-structure than does the 4 LFSK, as is seen in the spectra of Figures 47 through 50. All of the spectra have widths of approximately 4 divisions, or 20 kHz, where the width is determined 10 dB down from the peak. This effect is due to the peak deviation caused by all the waveforms being the same, as all the quasi-analog signals have the same amplitude. Thus their Carson's rule bandwidths will be nearly equal, as the baseband converter signal bandwidths are less than the radio frequency deviation for all these cases.

The X-mode FM spectra, shown in Figures 51 through 54, all show a lobing structure, similar to that of the quasi-

TABLE 7

FM SPECTRA MEASUREMENT PARAMETERS

Figure	Code	Bit Rate	Mod. Index	kHz/Div.
47	2-LFSK	1200	1	5
48	2-LFSK	1200	0.5	5
49	4-LFSK	1200	0.8(≡0.27)	5
50	4-LFSK	1200	1.9(≡0.63)	5
51	MAN	9600	-	10
52	PST	9600	-	10
53	TWB/DICODE	9600	-	10
54	DLM	9600	-	10

13

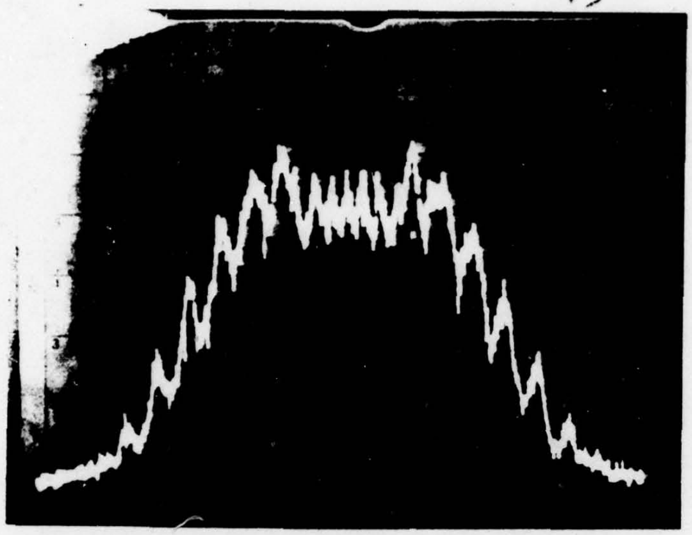


Fig. 47. **Binary FSK/FM Spectrum**
1200 bps, Mod. Index = 1.0,
10 dB/Div. Vert., 5 kHz/Div. Horiz.

14

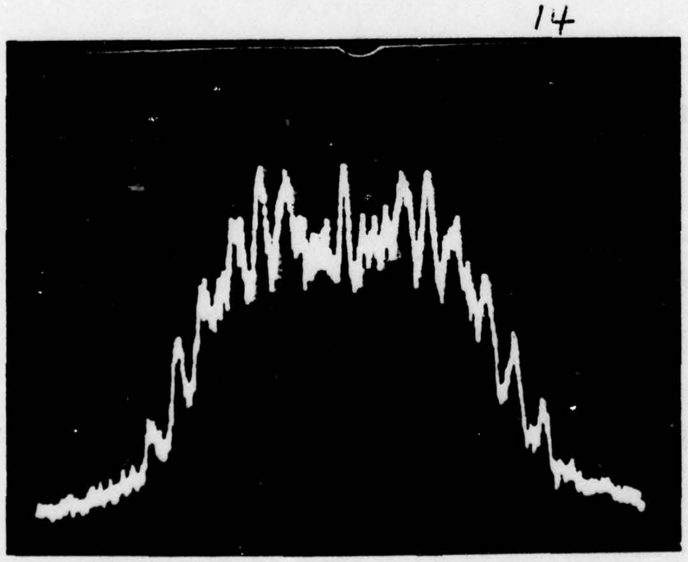


Fig. 48. **Binary FSK/FM Spectrum**
1200 bps, Mod. Index = 0.5,
10 dB/Div. Vert., 5 kHz/Div. Horiz.

15

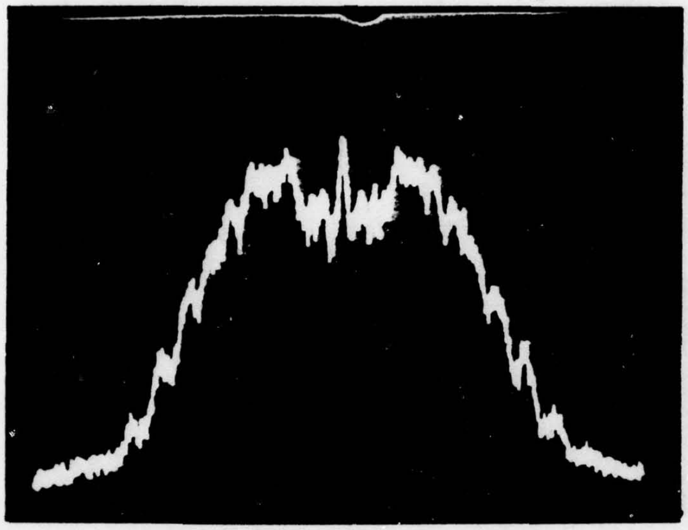


Fig. 49. 4 Level FSK/FM Spectrum
1200 bps, Mod. Index = 0.27,
10 dB/Div. Vert., 5 kHz/Div. Horiz.

16

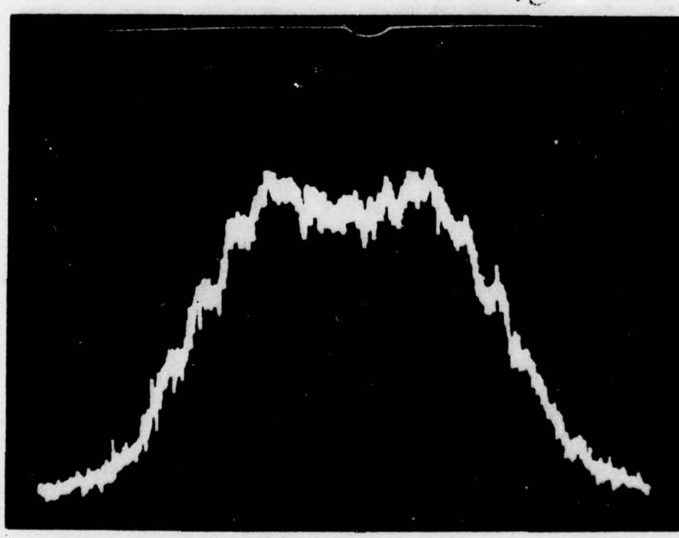
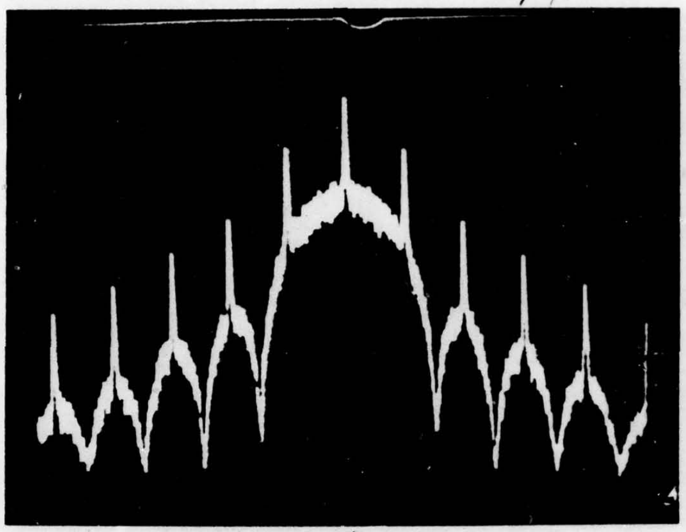


Fig. 50. 4 Level FSK/FM Spectrum
1200 bps, Mod Index = 0.63,
10 dB/Div. Vert., 5 kHz/Div. Horiz.

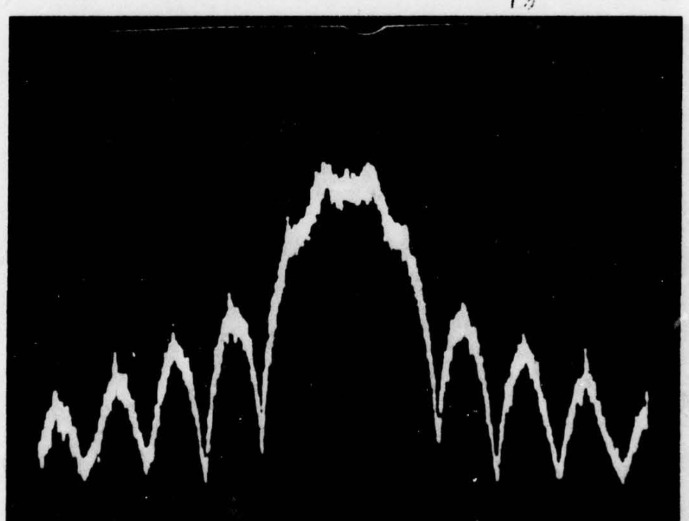
17



↑CARRIER

Fig. 51. Manchester/FM Spectrum
9600 bps, 10 dB/Div. Vert.,
10 kHz/Div. Horiz.

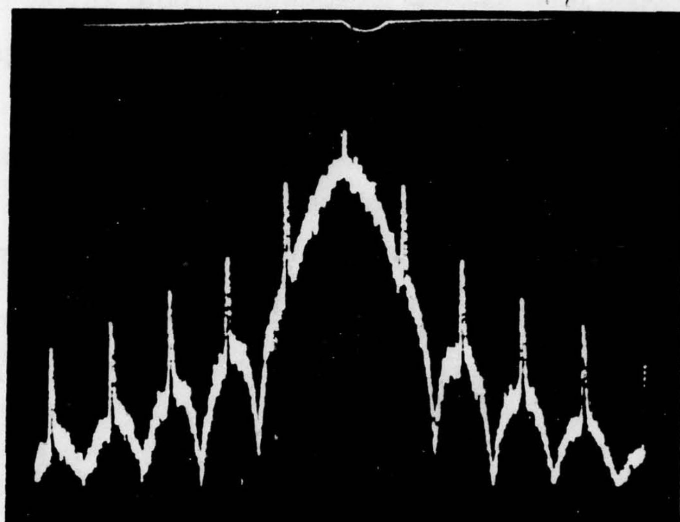
18



↑CARRIER

Fig. 52. Pair Selected Ternary/FM Spectrum
9600 bps, 10 dB/Div. Vert.,
10 kHz/Div. Horiz.

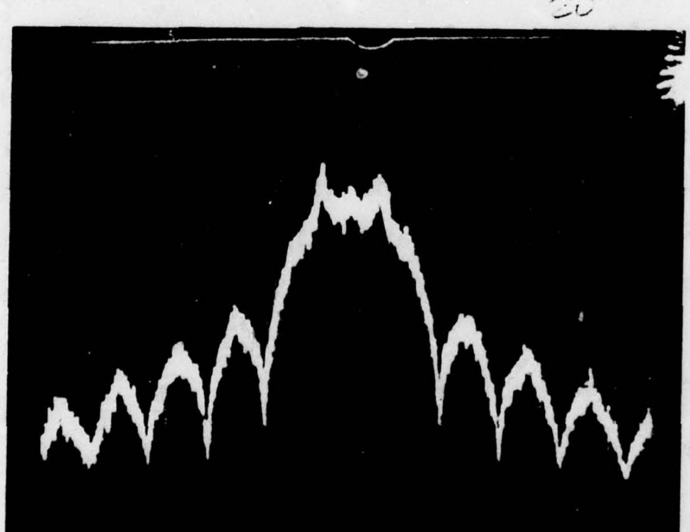
19



↑ CARRIER

Fig. 53. Twinned Binary/FM Spectrum
9600 bps, 10 dB/Div. Vert.,
10 kHz/Div. Horiz.

20



↑ CARRIER

Fig. 54. Delay Modulation/FM Spectrum
9600 bps, 10 dB/Div. Vert.,
10 kHz/Div. Horiz.

analog baseband signals. This is explained by the fact that both modulations are direct frequency modulation by digital data. The width of the main lobe of all of the spectra is three times the data rate, and the discrete lines, when present, are speced at the bit rate frequency. An examination of the X-mode FM spectra of Section 2.4.1 shows that the detailed spectra are a strong function of the data rate. This is understandable as the modulation index is, for a fixed frequency deviation, proportional to the data rate. Thus the spectra of Figures 51 through 54 are simply a sample at one data rate, 9600 bps. The computed spectra of Section 2.4.1 can be used as a guide to the effect of changing the X-mode data rate.

4.2 Error Rate Measurements

Figures 55 through 72 show the measured and theoretical error rates for all of the X-mode and quasi-analog mode signals. All measurements were made using two Baseband Signal Converters connected back-to-back. Additive Gaussian noise generated by a General Radio Model 1381 Random Noise Generator, was inserted into the receiving system, and its level was varied using a calibrated attenuator to change the signal-to-noise ratio. The data sequence used in the testing was a maximal length sequence, generated by a 24 stage shift register. The sequence length was therefore $2^{24}-1$ bits.

The order of the figures is as follows:

From Fig.	Through Fig.	Mode	SN Transmitter	SN Receiver
55	60	X	1	2
61	66	X	2	1
67	69	Q.A.	1	2
70	72	Q.A.	2	1

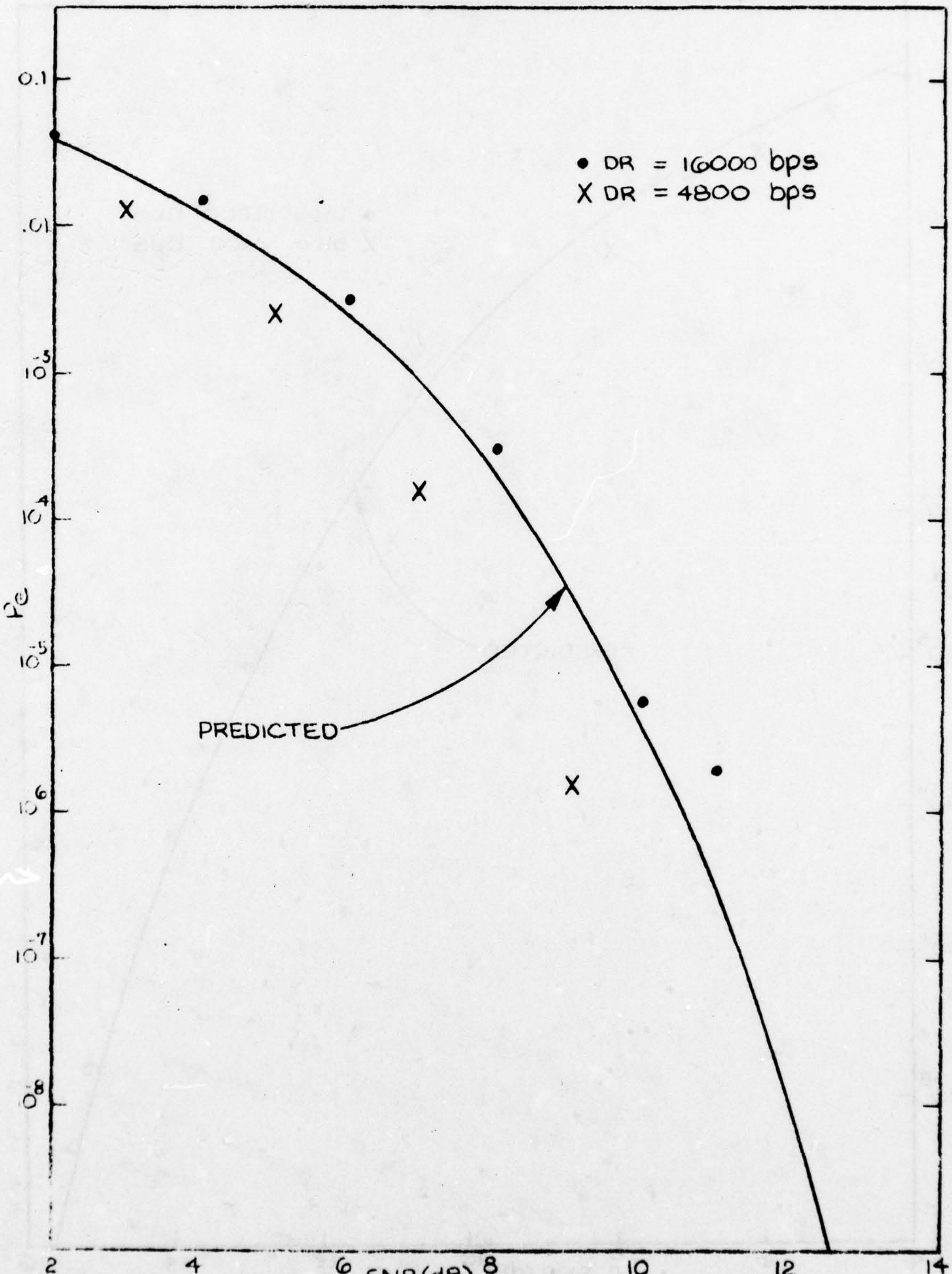


Fig. 55. Measured Manchester Code Error Rate. SN1 Driving SN2

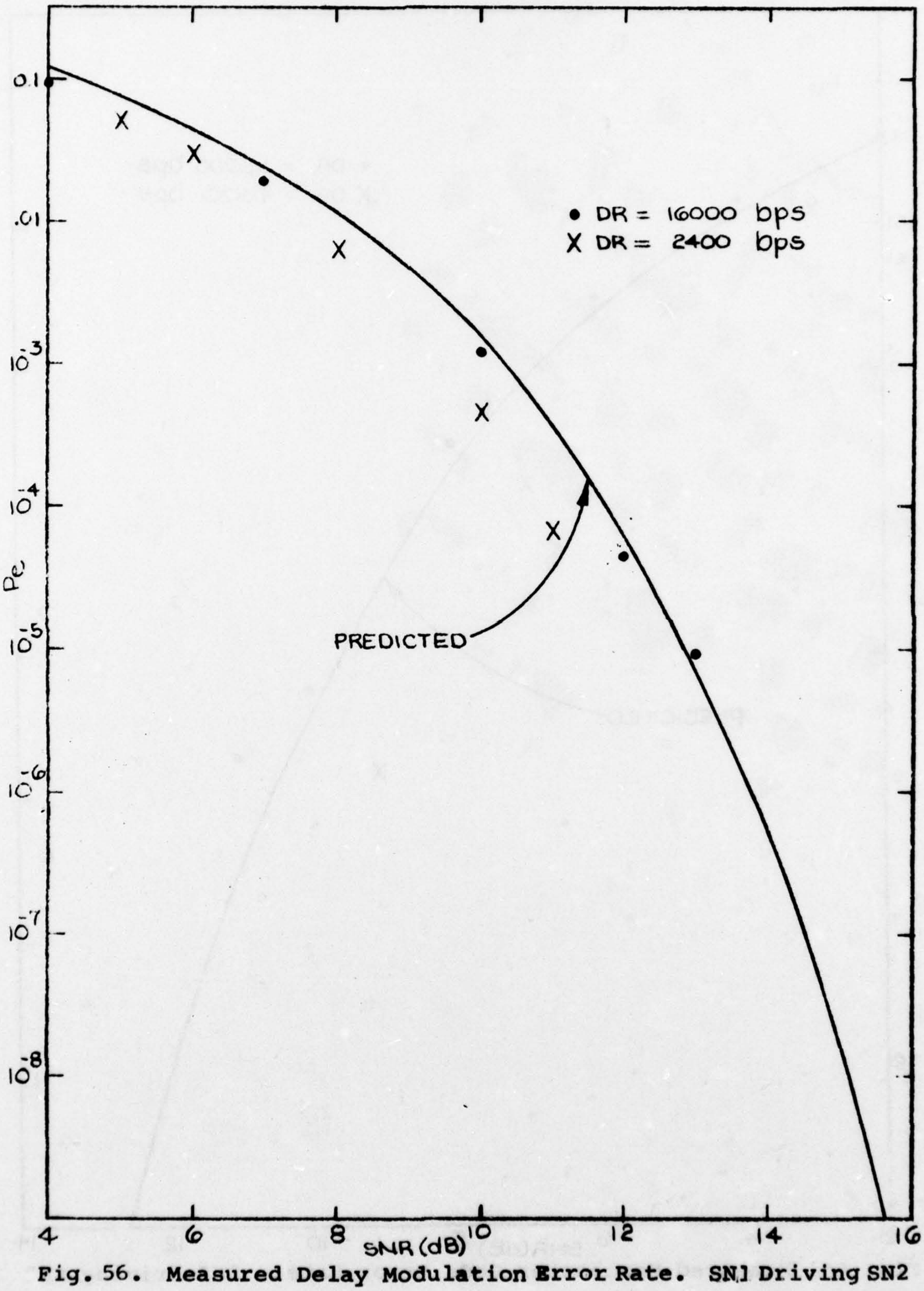


Fig. 56. Measured Delay Modulation Error Rate. SN1 Driving SN2

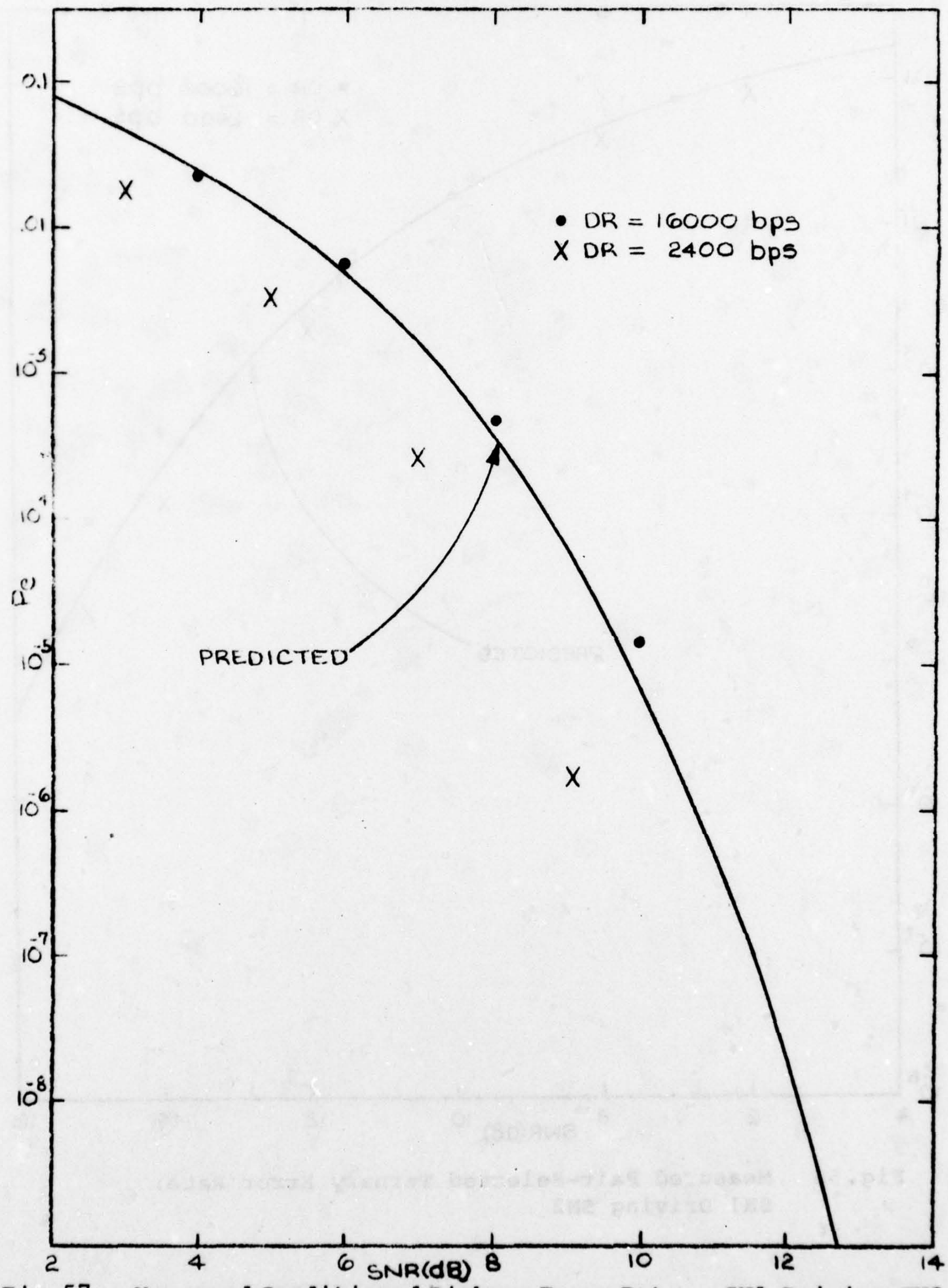


Fig.57 Measured Conditioned Diphase Error Rate. SN1 Driving SN2

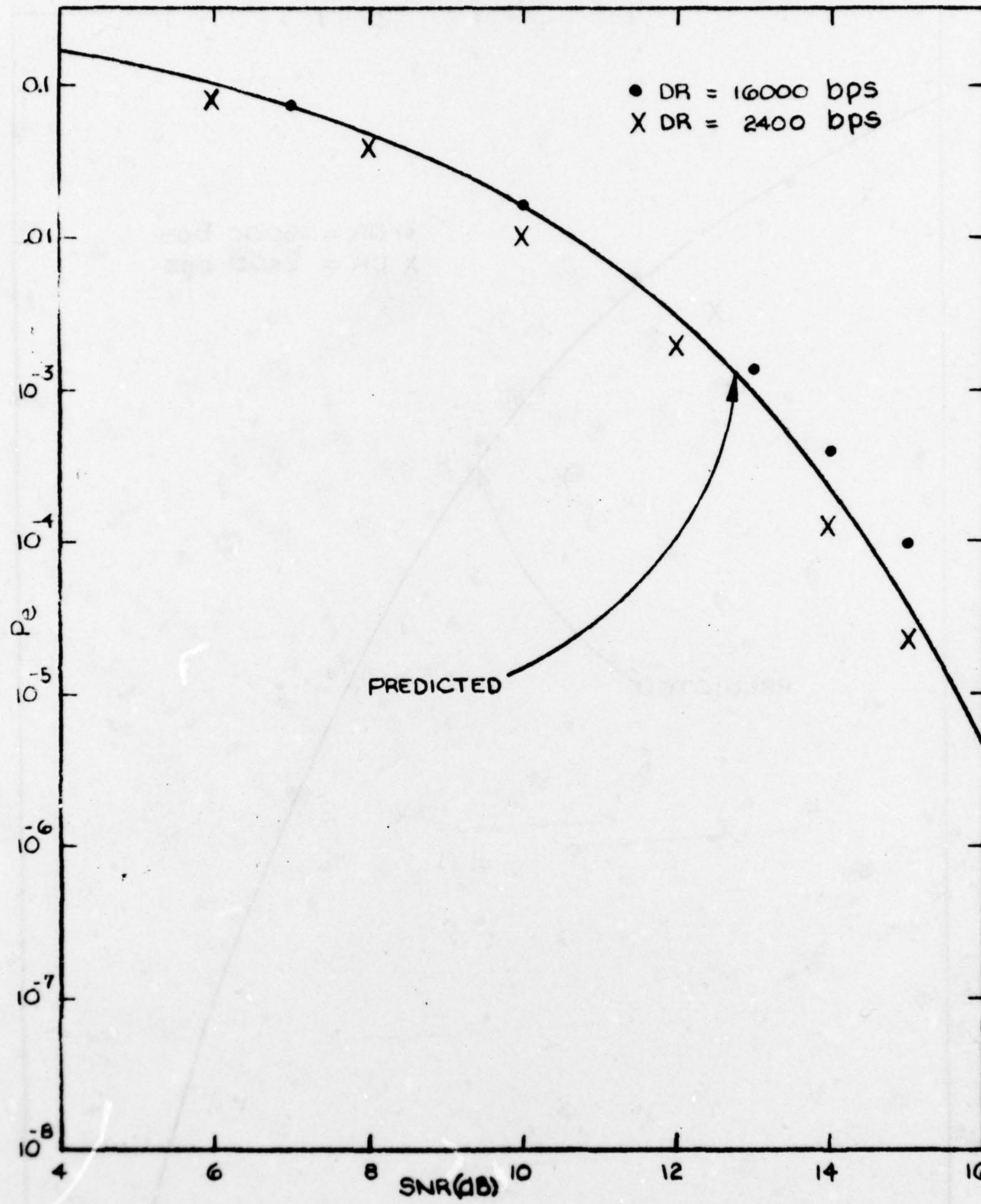


Fig. 58 Measured Pair-Selected Ternary Error Rate. SN1 Driving SN2

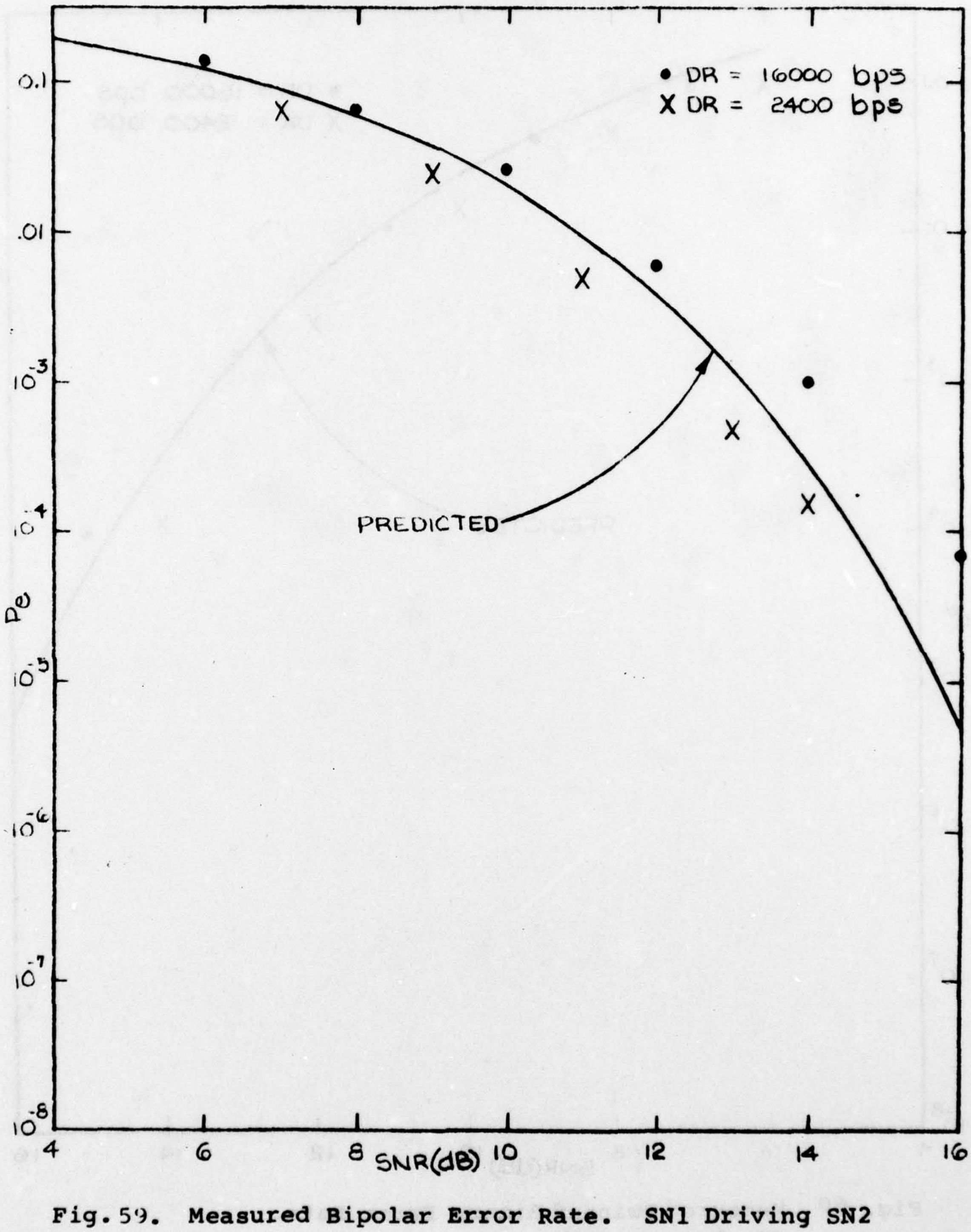


Fig. 59. Measured Bipolar Error Rate. SN1 Driving SN2

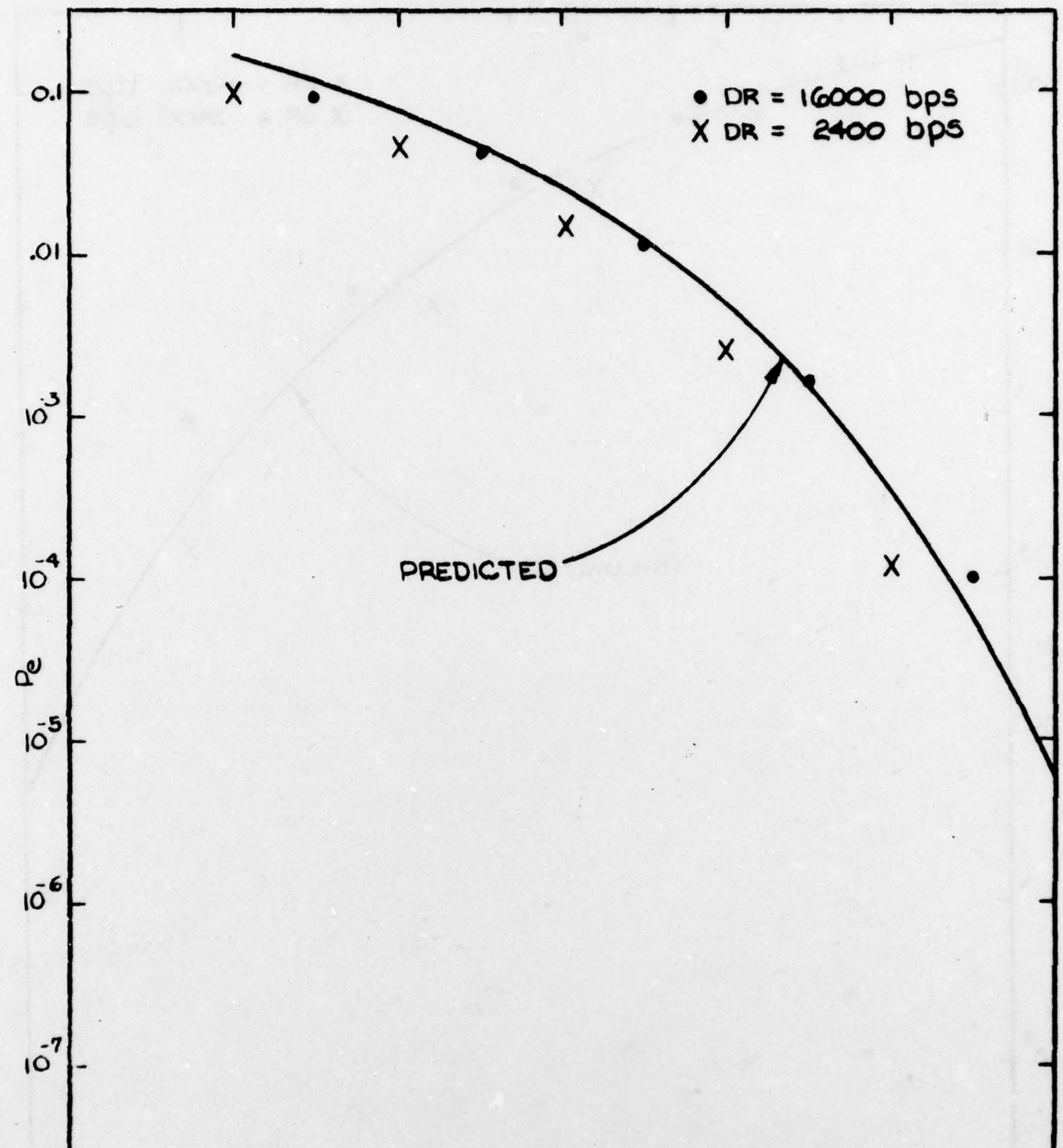


Fig. 60 Measured Twinned Binary Error Rate.
SN1 Driving SN2

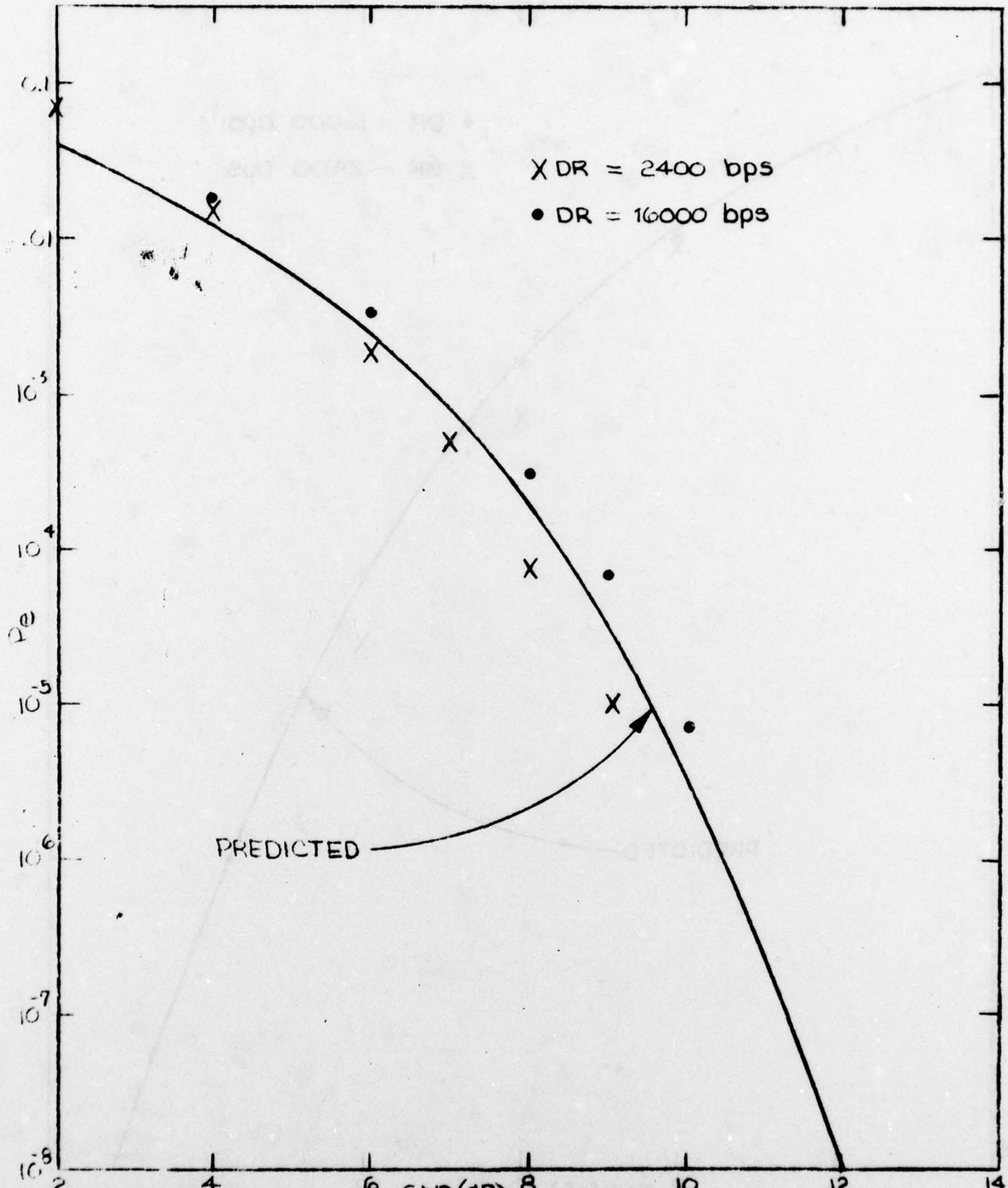


Fig. 61 Measured Manchester Code Error Rate. SN2 Driving SN1

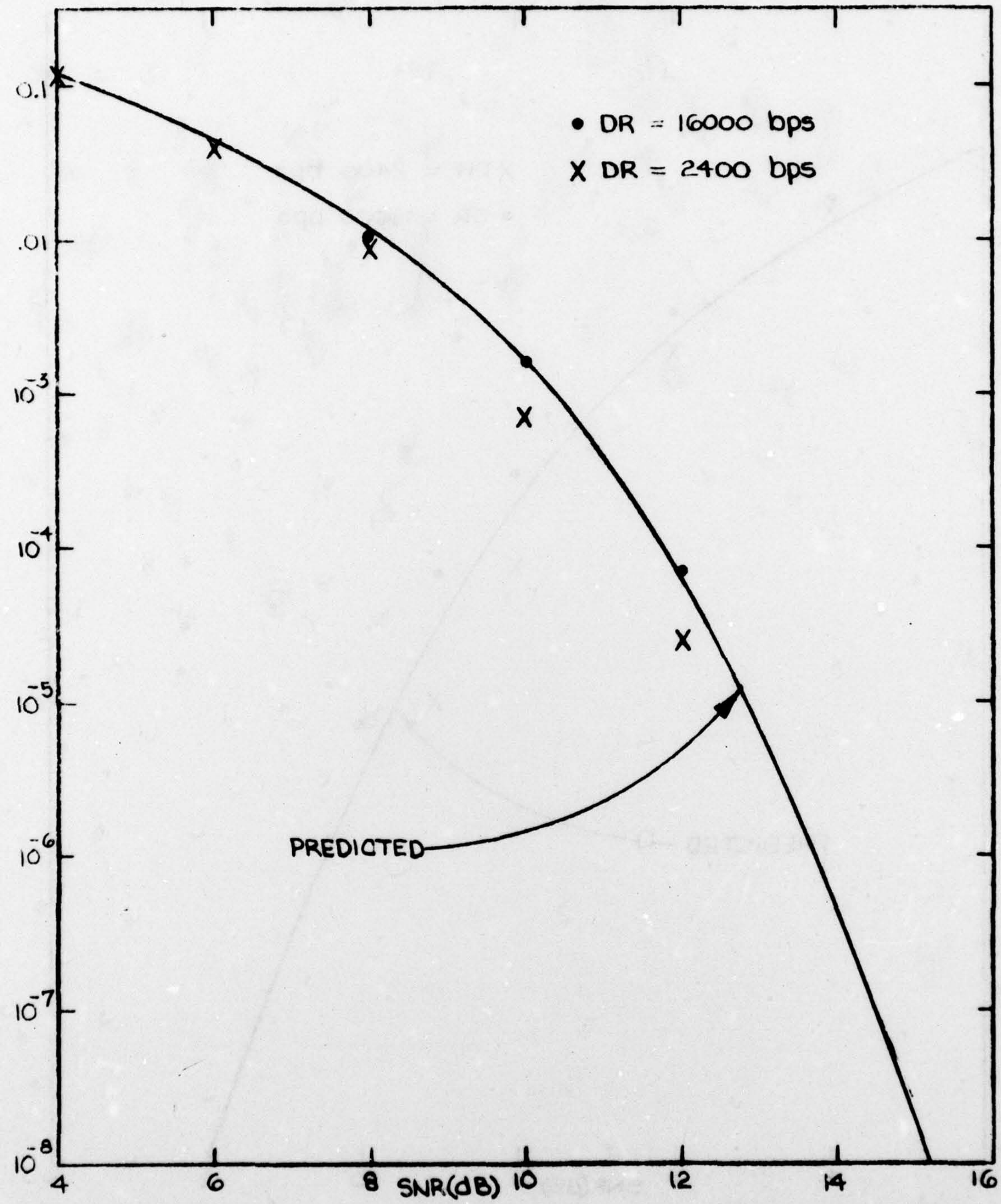


Fig. 62 Measured Delay Modulation Error Rate. SN2 Driving SN1

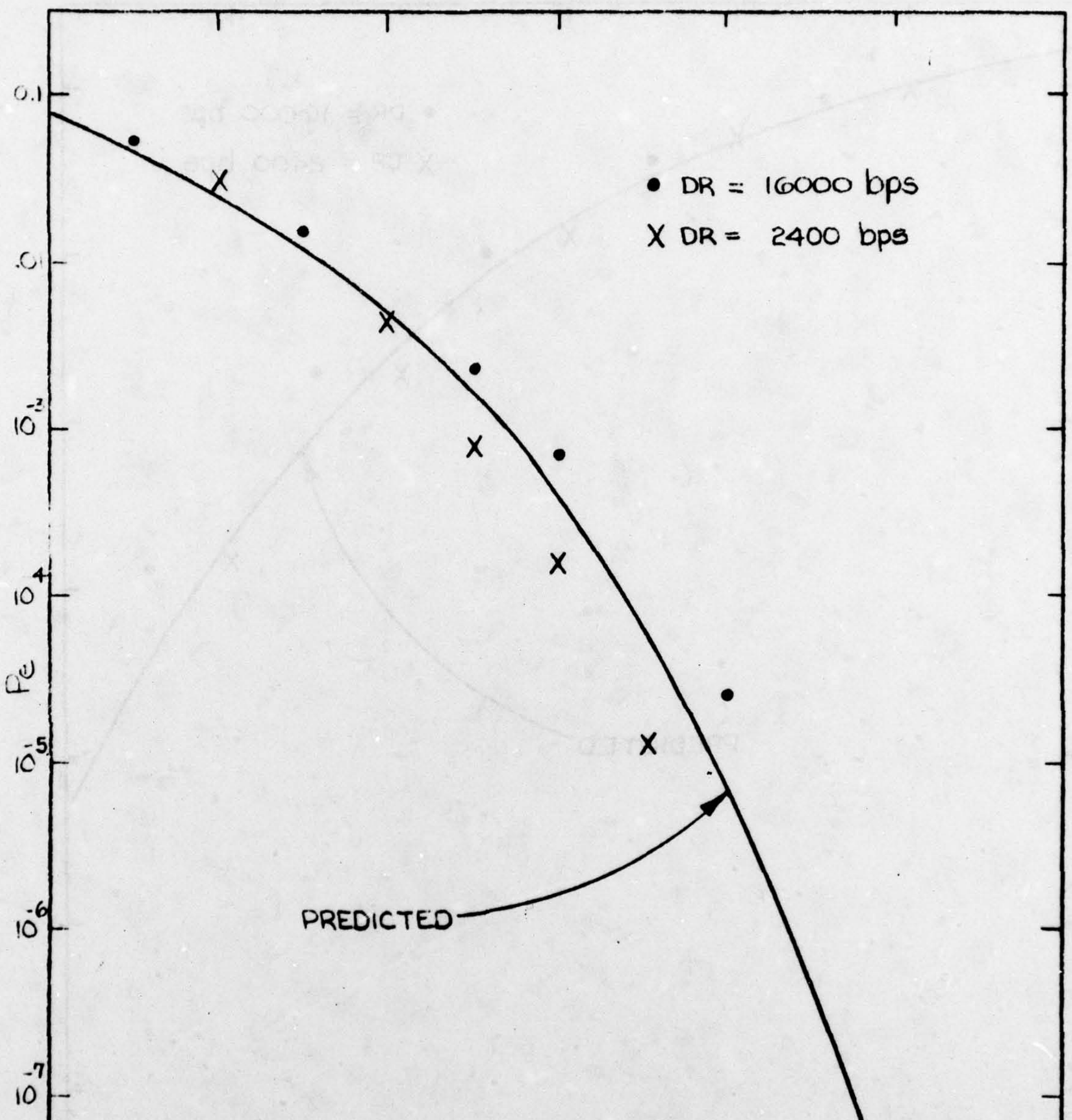


Fig. 63 Measured Conditioned Diphase Error Rate. SN2 Driving SN1

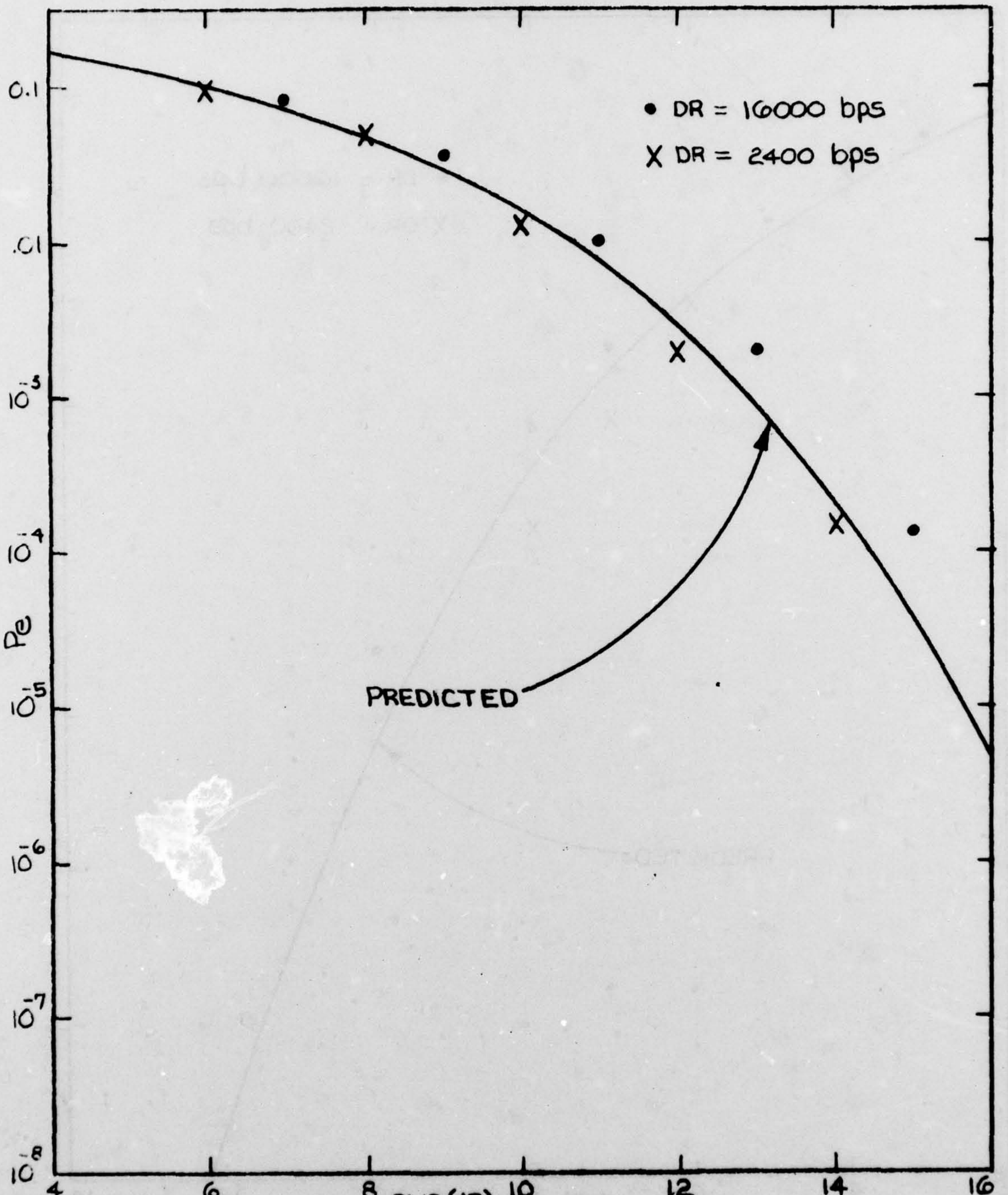


Fig. 64 Measured Pair-Selected Ternary Error Rate. SN2 Driving SN1

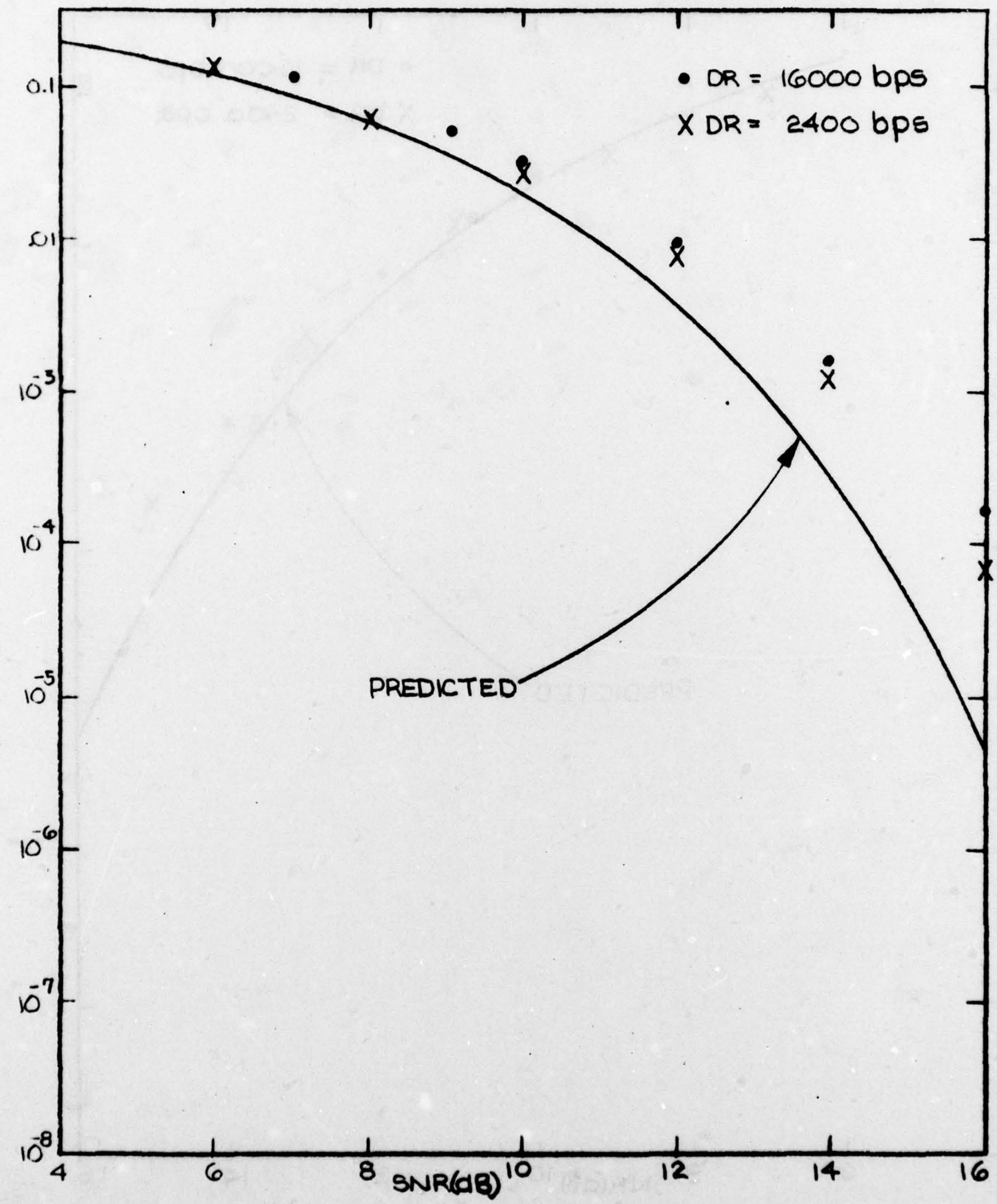


Fig. 65 Measured Bipolar Error Rate. SN2 Driving SN1

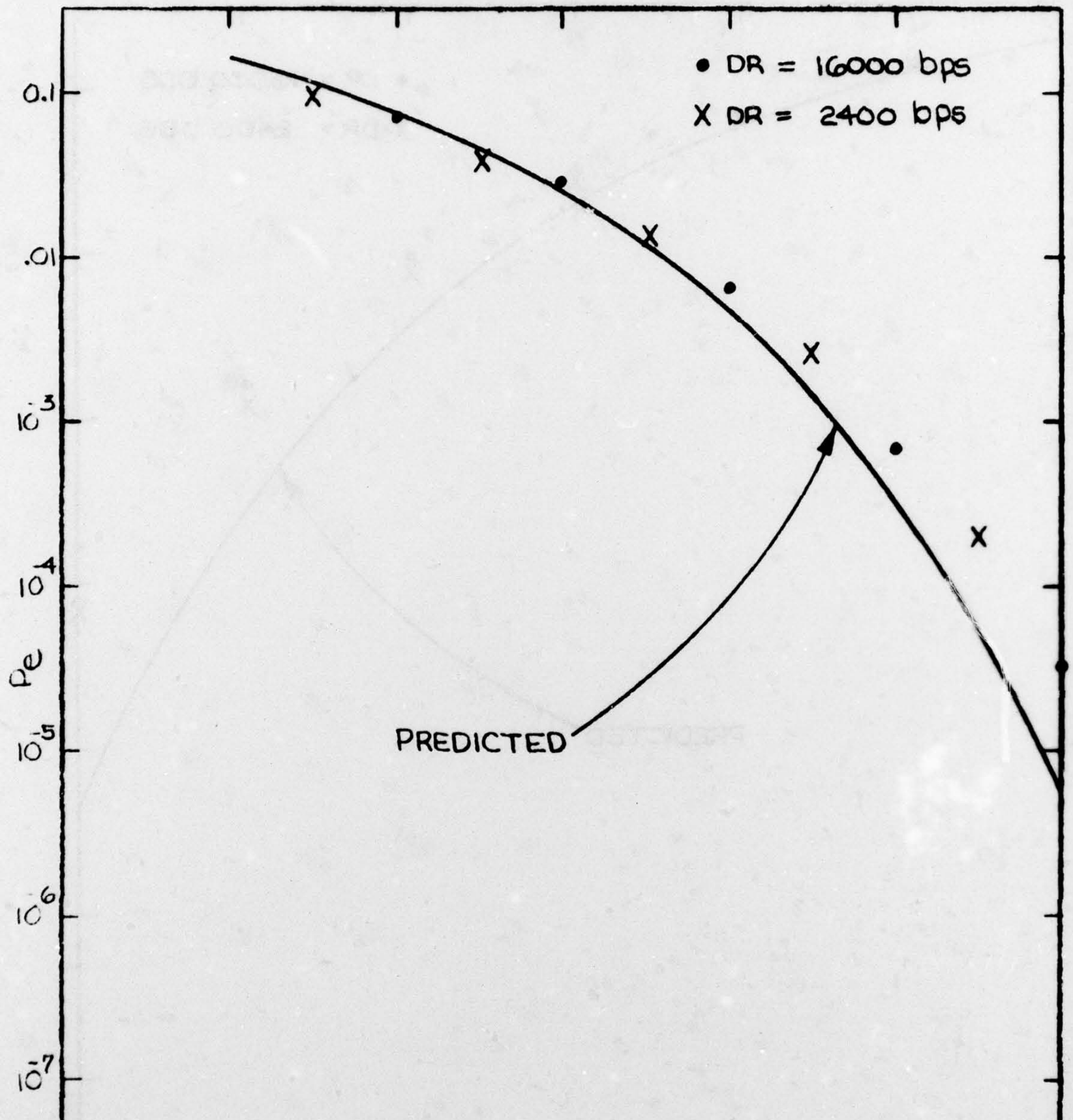


Fig. 66 Measured Twinned Binary Error Rate.. SN2 Driving SN1

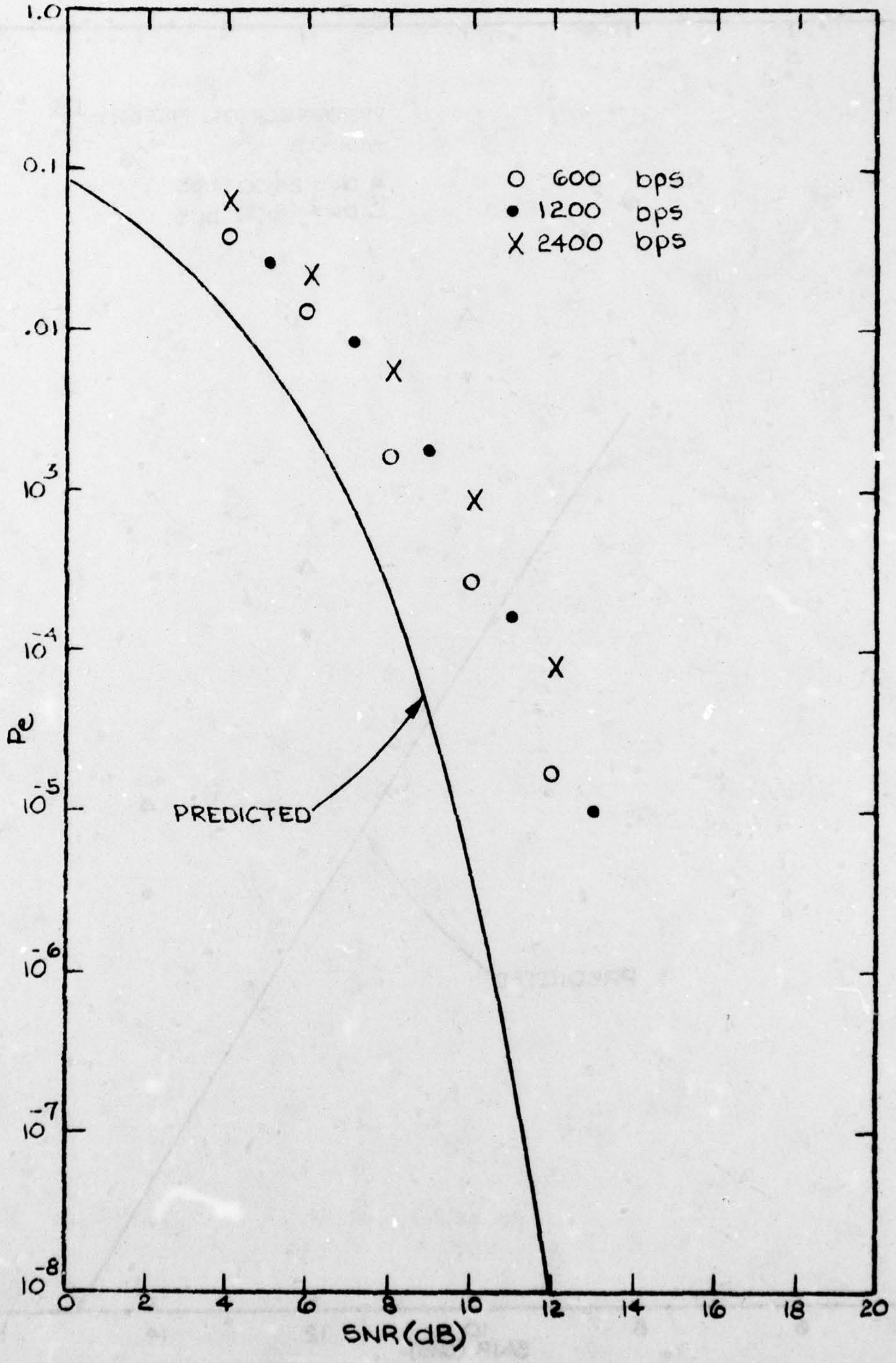


Fig. 67 Measured MSK Error Rate. SN1 Driving SN2

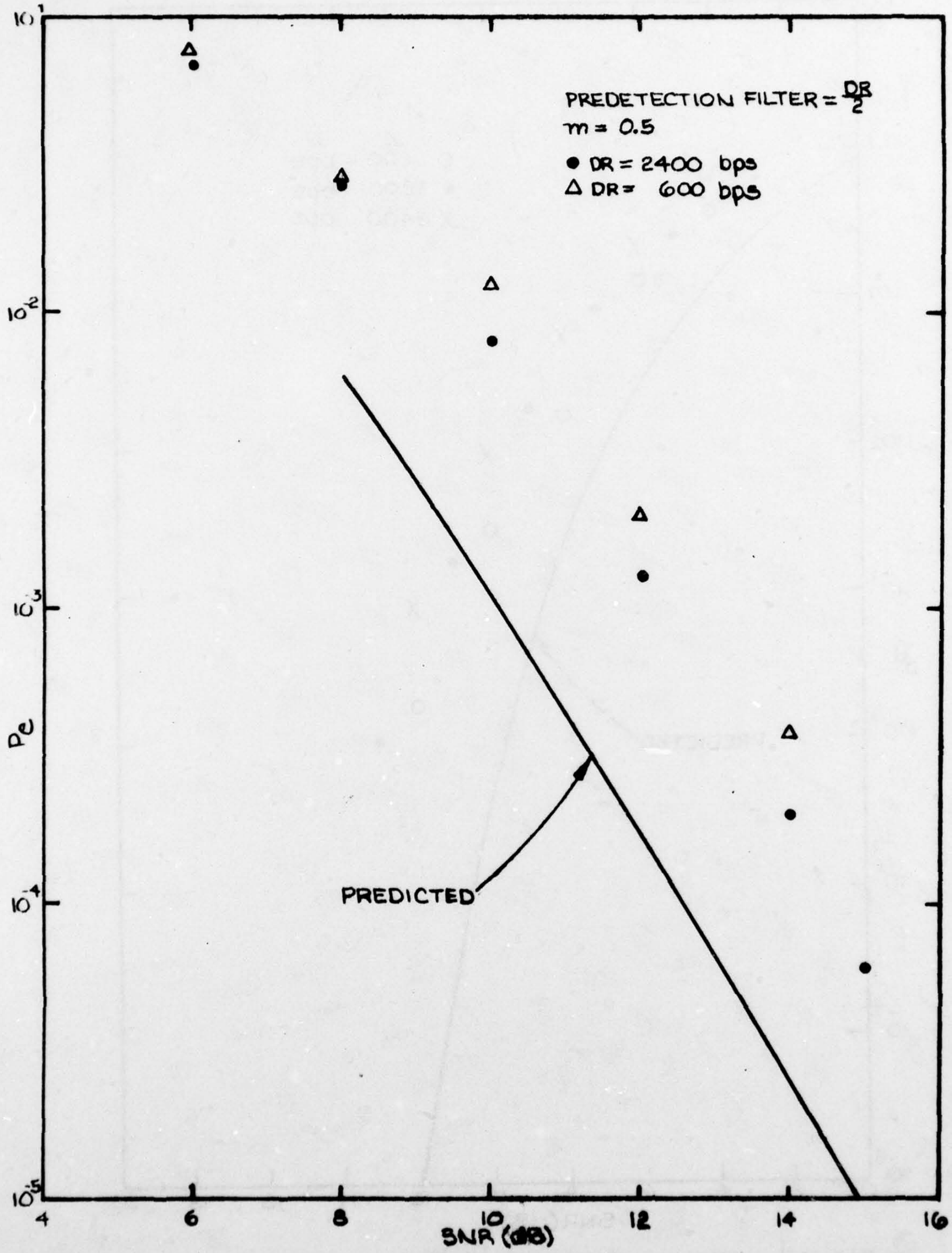


Fig. 68 **Measured Binary FSK Error Rate. SN1 Driving SN2**

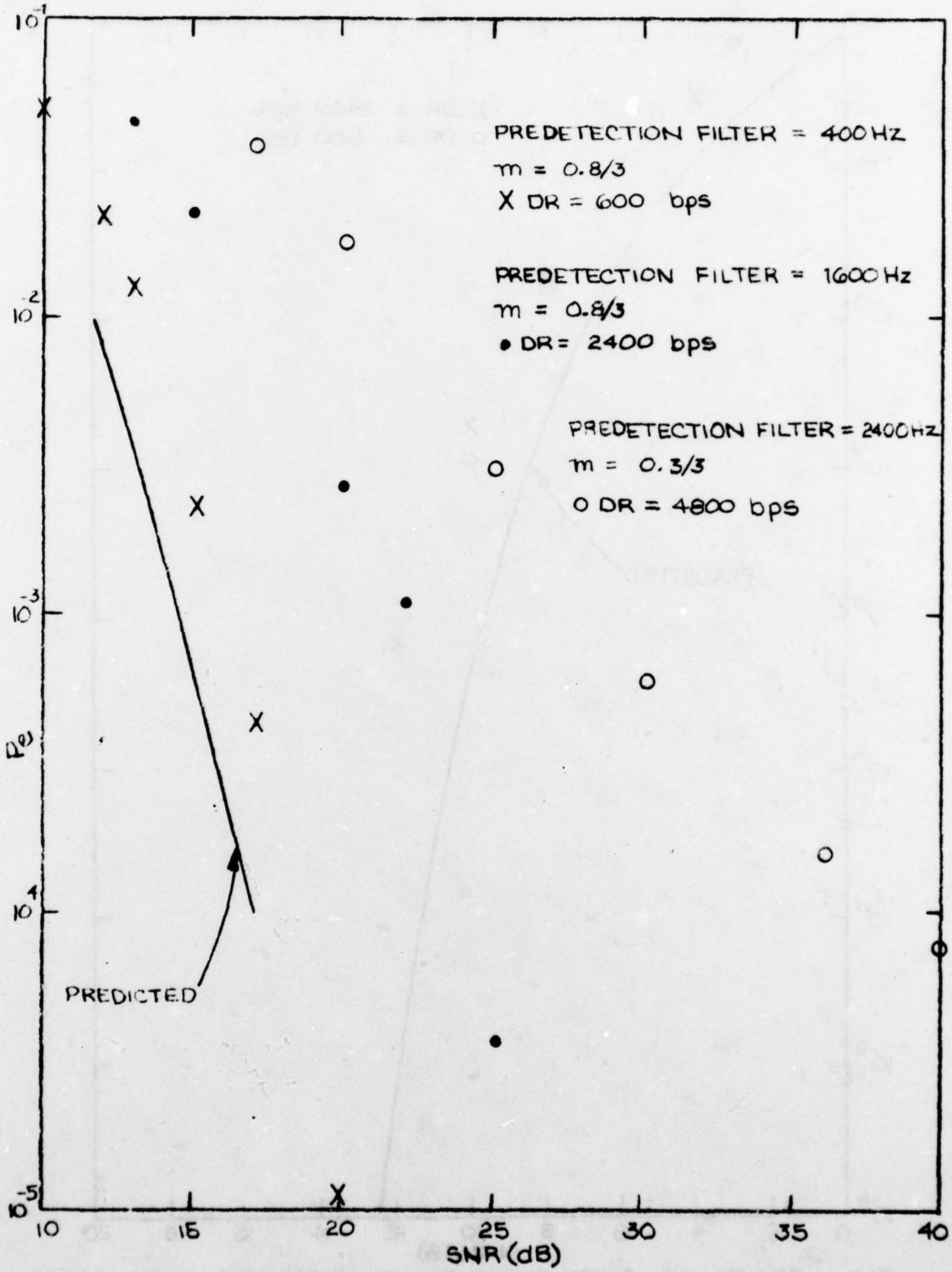


Fig. 59 Measured 4LFSK Error Rate. SN1 Driving SN2

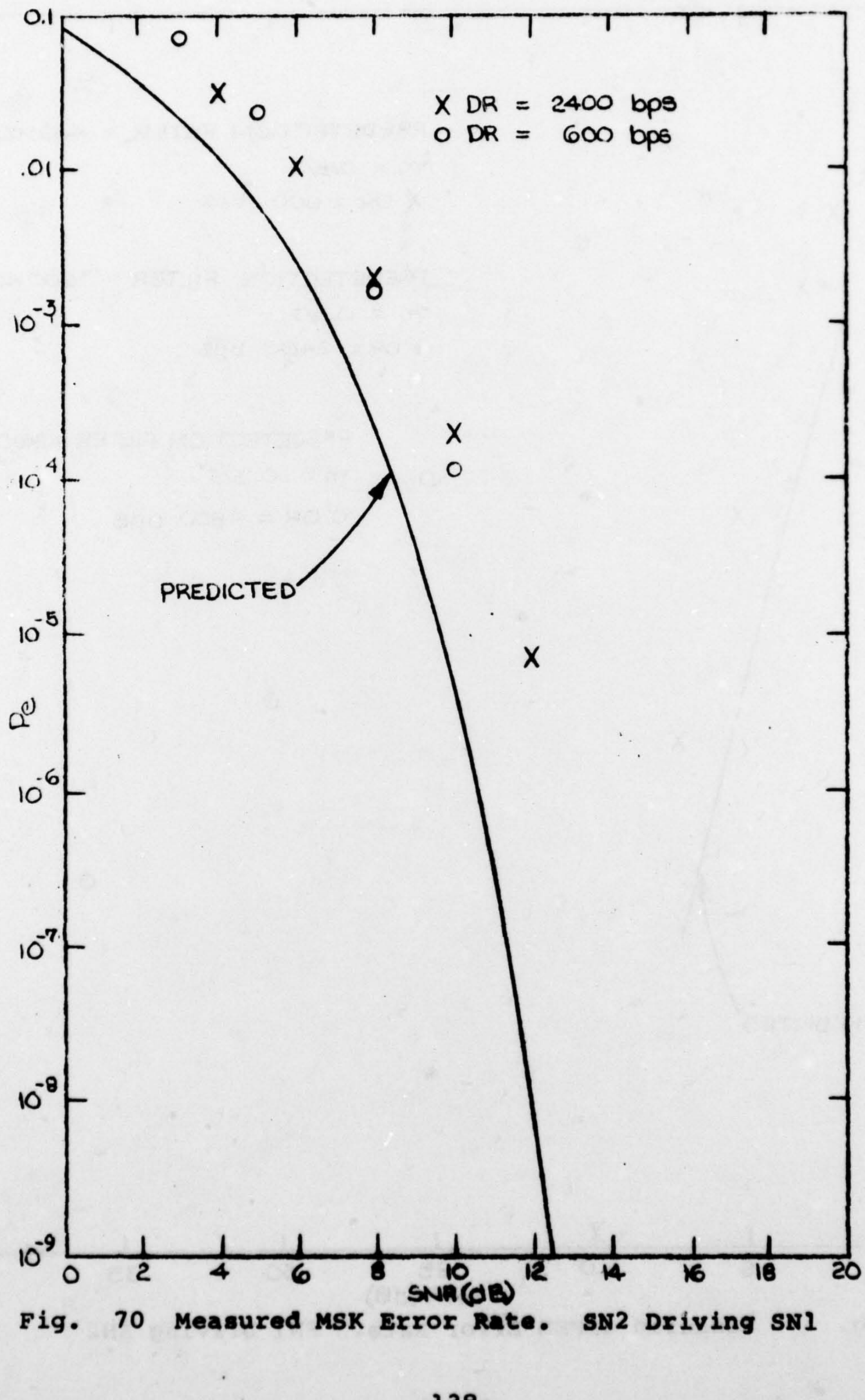


Fig. 70 Measured MSK Error Rate. SN2 Driving SN1

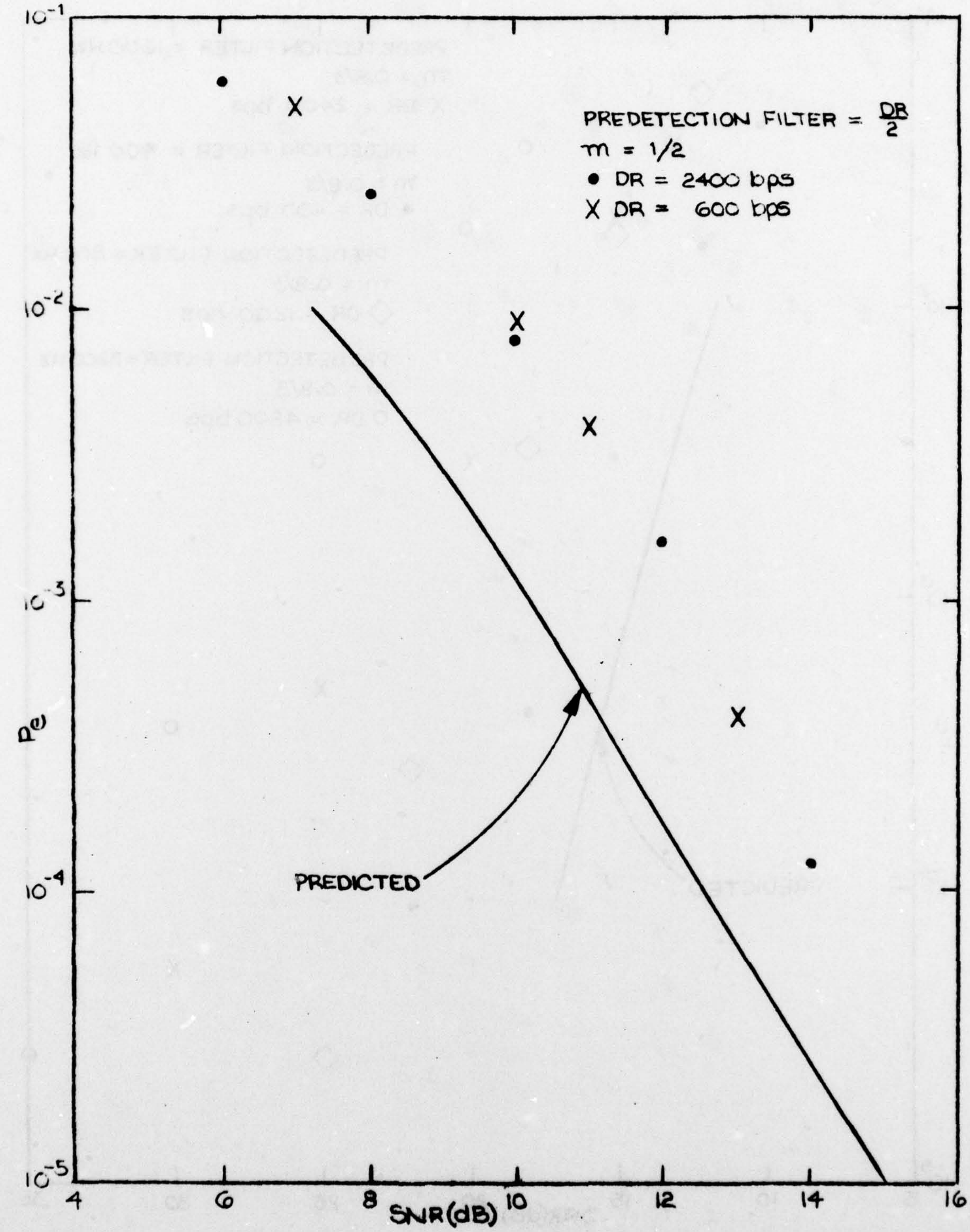


Fig. 71 Measured Binary FSK Error Rate. SN2 Driving SN1

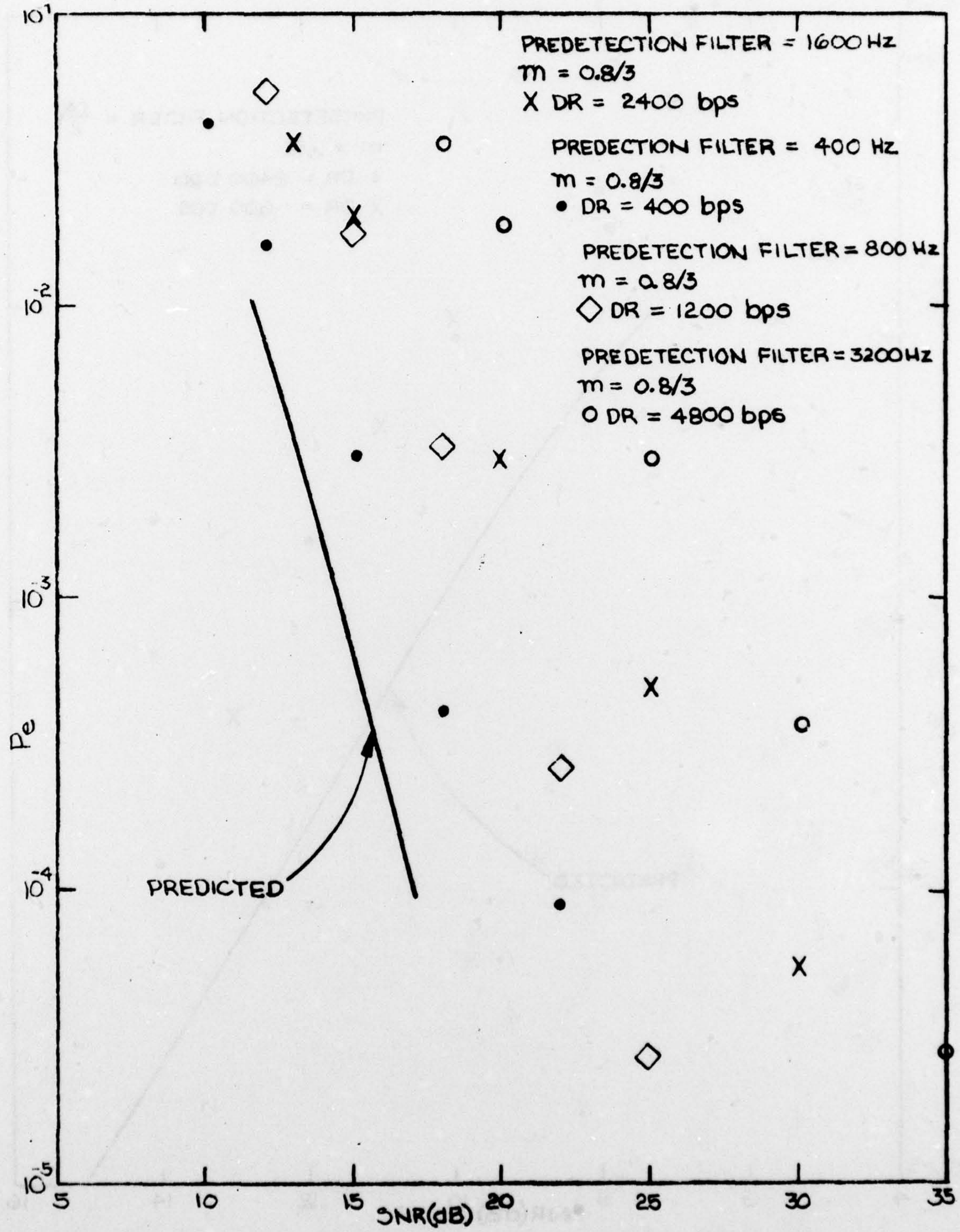


Fig. 72 Measured 4LFSK Error Rate. SN2 Driving SN1

Each Figure contains error rate curves as a function of signal-to-noise ratio, taken at two or more data rates. The experimental curves can be summarized as follows:

(1) The measured X-mode error rates are in excellent agreement with the predicted values. There is little data rate dependence, and the implementation loss appears to be negligible.

(2) The measured Quasi-Analog mode error rates show an implementation loss of about 2 dB for MSK and binary FSK, with little data rate dependence. The implementation loss of 4 LFSK is data rate dependent, being about 2 dB at 600 bps and 7 dB at 2400 bps, when the modulation index is 0.8/3, or 0.27. In order to operate at 4800 bps it is necessary to decrease the modulation to 0.1, which results in a signal-to-noise ratio degradation of about 15 dB.

(3) There is no significant performance difference between the two delivered units.